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A RECEIVER FOR THE RADIO WAVES FROM INTERSTELLAR HYDROGEN

I. THE INVESTIGATION OF THE HYDROGEN RADIATION

by C. A. MULLER *).

522.6:621.396.9

Radio astronomy, born just before the second world war, has developed with amazing rapidity. Radio astronomical research can now boast of results of far-reaching importance, which have considerably improved and sharpened our picture of the universe. The instruments employed in radio astronomy are for the greater part electronic. In the Netherlands, the author, member of the Netherlands Foundation for Radio Astronomy, has headed the development of a receiving installation for research on the radio spectral line of wavelength of 21 cm, which is radiated by the hydrogen atoms in interstellar space. This receiver has been used in conjunction with a 7.5 m diameter parabolic reflector at Kootwijk. A similar receiver is destined for a "radio observatory" at Dwingelo. The receiver will be described in two articles; the first, printed below, outlines the place of radio astronomy in modern astronomical research and discusses the nature and properties of the radiation and the requirements which the receiver must fulfil. The second article will give a detailed description of the receiver circuits.

Radio waves from space

One of the most important discoveries in astronomy in this century was the observation of radio waves originating outside the Earth. This radiation, discovered by chance in 1931 by Jansky in the United States, was first systematically investigated a number of years later by the amateur Reber. Not until during the second world war was work in this field begun in earnest, first in Australia and in England. At present, radio astronomy is carried on in a number of institutes spread over the whole world.

The reception of the radio waves requires, of course, instruments which are utterly different from the usual optical instruments used in conventional astronomy. Fortuitously, after the second world war, many radar installations for which there was now no further use could be found in various countries. These were very well suited for use as receivers for the V.H.F. radio waves from space. In the Netherlands, an abandoned German radar antenna of 7.5 m diameter was moved by the P.T.T. (Dutch Post Office) to Kootwijk Radio Station and later placed temporarily at the disposal of the Netherlands Foundation for Radio Astronomy. This body includes representatives from the Dutch Post Office, the observatories of Leiden, Utrecht and Groningen, and those of Belgium, the K.N.M.I. (Royal Dutch Meteorological Institute) and the Research Laboratory of N.V. Philips. The work of the Foundation is subsidised by the Netherlands Organisation for Pure Research (Z.W.O.). A group of investigators of the Foundation has been making continuous observations at Kootwijk, the results being worked out at the Leiden observatory. A new parabolic reflector of 25 m diameter has been built in the neighbourhood of Dwingelo and will shortly be brought into use. This antenna is movable in all directions, and has an automatic steering mechanism. In addition, two more parabolic antennae 7.5 m in diameter are being set up in Dwingelo.

Not all the radiation from space can reach the Earth: the atmosphere is transparent for wavelengths between about 1 cm and 20 m. Radiation

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with a wavelength shorter than 1 cm or longer than 20 m is almost completely absorbed in the atmosphere or reflected from the ionosphere. The wavelength range of this "radio window" is, however, very wide (8 to 10 octaves) compared with the narrow "optical window": the atmosphere of the Earth transmits "by chance" just the visible range of the spectrum, but this covers less than one octave (3700 Å — 8000 Å). Here we see one of the reasons for the importance of radio astronomy.

It was soon found that the intensity of the radiation differed for the various wavelengths within the transmitted range: we can thus speak of a radio spectrum. The optical spectra of stars, e.g. that of the sun, exhibit a great number of lines (corresponding to atomic transitions) against a background of the continuous emission spectrum (caused mainly by the capture of electrons by hydrogen atoms). The radio spectrum is probably mainly a continuous emission spectrum; discrete atomic or molecular transitions corresponding to frequencies in the radio range have a very small probability. In general, therefore, no spectral lines can be observed here on Earth. However, in 1944 the Dutch astronomer Van de Hulst [1]) drew attention to the possibility that at least one emission line in the radio range, viz. at wavelength of 21 cm, should be observable. Radiation of this wavelength is emitted by the highly rarefied clouds of hydrogen between the stars and which, together with the equally rarefied dust clouds, constitute the matter of interstellar space.

As a result of the extreme rarefaction, the hydrogen in interstellar space can exist in atomic or ionized form. Until recently, one could observe hydrogen gas only in the neighbourhood of very hot stars, where it is almost completely ionized. One of the effects of the capture of electrons by hydrogen ions is the emission of visible spectral lines and a continuum. Only indirect methods had been available by which anything could be concluded about the physical condition of the more extended regions of the neutral hydrogen. Van de Hulst argued that for the transition in the hydrogen atom giving rise to a quantum of wavelength 21 cm, there was a good chance that it would be observable on Earth. In fact, it is possible to observe this spectral line. The calculation involved in this argument is a good illustration of the magnitudes juggled with by astronomers. Any arbitrary H-atom exhibits this transition only about once in 11 million years; furthermore the density of the interstellar hydrogen is extremely low, of the order of 1 atom per cm^3 (1 cm^3 hydrogen at 1 atm pressure contains at room temperature about 2.5×10^{19} atoms). However, since the hydrogen clouds stretch out over enormous regions — they are spread throughout the whole of the Galactic system, with an equatorial diameter of 10^{23} cm the power intercepted on Earth from particular directions is still just enough to observe.

An extensive search was started for this spectral line, and in 1951 it was indeed found, almost simultaneously by workers in the United States, in Holland and in Australia [2]).

The full importance of this discovery can hardly be overestimated: a new tool was now available for investigating the structure of the Galactic System, a tool which was much more direct than those hitherto available. We will return to this in the next section.

The intensity of the hydrogen radiation is very weak; the power received is many times (10 to 1000 ×) less than the noise level which is produced in the receiver itself. The intercepted radiation also has the character of noise, like the receiver noise. The value of the intensity is a function of the frequency, with a marked maximum around 1420 Mc/s, corresponding to the wavelength 21.1 cm.

The noise character of the interstellar radio waves, i.e. the statistical variations in their intensity with time, is not to be unexpected. It is caused by the fact that the sources of radiation, i.e. the hydrogen atoms, emit their radiation independently of each other. The number of atoms in which the transition occurs in unit time, and which therefore send out a quantum of energy, fluctuates statistically with time: the power intercepted on Earth therefore fluctuates too.

In optical astronomy, the receiving apparatus (e.g. a photocell) generally produces comparatively little noise, at most about as much as the noise power of the radiation itself. Use is often made of a photographic plate; the long exposure completely eliminates the effect of the noise. In radio astronomy the intensity of the received radiation, particularly for short waves, is small with respect to the receiver noise. This is also true for the 21.1 cm radiation; hence the receiver noise has a critical effect on the accuracy of the measurement. In principle, the effect of the receiver noise can be reduced by means of long integration times, which imply longer periods

[1]) H. C. van de Hulst, Ned. T. Natuurk. **11**, 201, 1945.

[2]) H. I. Ewen and G. M. Purcell; C. A. Muller and J. H. Oort; J. L. Pawsey; all in Nature **168**, 356-358, 1951. The amplifier then used at Kootwijk was developed in co-operation with Dr. F. L. H. M. Stumpers of the Philips Research Laboratory at Eindhoven.

for a given measurement. In practice, lengthening integration times is restricted by the unavoidable instability of the receiver. It is therefore necessary to take more sophisticated measures to reduce or eliminate its effect.

The Netherlands Foundation for Radio Astronomy, and in particular the Kootwijk group, has developed a receiving and amplifying installation which aims at squeezing out the last drop of available information. This is the receiver to be described in these pages. Before embarking on the description it may be useful to say a few words about some of the results so far obtained in the study of the hydrogen radiation, to give an impression of the many achievements of this new method and its potentialities.

Radio-astronomical observations at a wavelength of 21 cm

As already stated, observations of the interstellar hydrogen line yield information on the structure of the Galactic System. The latter consists of a great number of stars (about 10^{11}), distributed in space roughly in the form of a flattened disc with a diameter of about 30 000 parsec (1 parsec $\approx$ 3 light years $= 3 \times 10^{18}$ cm), and an average thickness of about 1000 parsec (*fig. 1*). This disc rotates on an axis through its centre, perpendicular to the galactic plane, the rotation time in the neighbourhood of the sun being about 200 million years. The galaxy is not composed only of stars: there is also interstellar matter; the two are present in about equal quantities as far as mass is concerned. The density of the material in the galaxy is not homogeneous. Stars frequently occur in groups, whose members can be distinguished from other stars in that they all perform nearly enough the same motion; also the interstellar matter is concentrated in clouds. Certain galactic bodies have preferred positions near the equatorial plane; others, on the other hand, are only to be observed at high "galactic latitude". Gas and dust clouds, for example, are always to be found in the region of the equator. This is a rather a pity, since these clouds with their light-scattering properties obscure a large section of the galaxy from us. The solar system is also situated near the equatorial plane and at a great distance from the centre. The absorbing matter makes it quite impossible to observe the region around the centre optically; it is only possible to make indirect deductions about this region from observations of objects in the neighbourhood of the Sun and from observations at higher galactic latitude, in which we look past the centre. It is therefore not possible to determine accurately the direction and distance of the centre. Clearly then, the information obtainable about the Galactic System by optical means is necessarily vague. The first detailed results come from very recent work and of necessity concern only the immediate neighbourhood of the Sun. They lead to some indications of the existence of a spiral structure, such as occurs in many galaxies outside our own, but whose existence in our own galaxy has not hitherto been demonstrable.

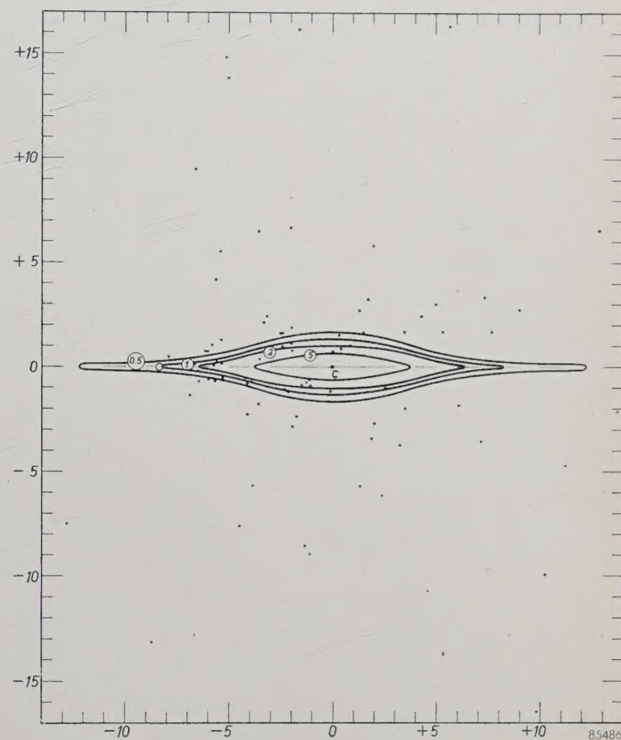

Fig. 1. Schematic section of the Galactic System, perpendicular to the equatorial plane. The Sun is indicated by the open circle (between 0.5 and 1), the centre is the point *C*. The interstellar matter is shown as a narrow hatched band; the full lines are contours of equal density (relative values indicated by the encircled numbers; the contour containing the Sun is arbitrarily chosen as of unit density). The galaxy rotates about an axis through *C*, perpendicular to the equatorial plane. Scale: 1 unit = 1000 parsec.

Radio waves suffer much less from absorption by the interstellar matter than light. Interstellar gas, in general, absorbs very little but the interstellar dust, though having little effect on radio waves strongly scatters visible light. The cause of this lies in the fact that the diameter of the dust particles is of the same order as the wavelength of light. The radio waves are much longer and therefore suffer little scattering. Thus it is possible to observe sources of radio waves which are situated almost on the opposite side of the galaxy from the Earth.

From observations of other galaxies such as, for example, the spiral nebula in Andromeda, it seems that the interstellar matter occurs preferentially in the arms of the spiral. In our own galaxy, this is probably also the case. The concentration of material near the equatorial plane suggests this. Owing to the small absorption, which makes it possible to observe the hydrogen radiation from great distances, it seems possible to obtain more exact information about the structure of the galaxy as a whole from study of this radiation. To this end, an extensive mapping of the intensity and of the other characteristic properties of the 21 cm radiation over the whole sky is required. The results so far obtained at Kootwijk, give an impression of the distribution of the hydrogen clouds over the different directions in the galaxy as seen from the Sun. This, however, is not sufficient; it is also required to know the distance of the source of radiation coming from a given direction. Oort and co-workers [3]) have used an elegant method for determining this distance. This is based on the differential galactic rotation, the phenomenon that the angular velocity of rotation of the galaxy decreases from the centre outwards. This means that a hydrogen cloud viewed in a given direction, possesses a velocity component towards or away from the Earth purely as a result of the differential rotation. In *fig. 2* the

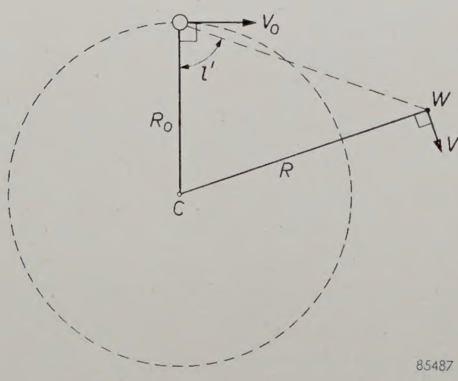

Fig. 2. Illustrating the relation between differential galactic rotation and radial velocity relative to the earth.

situation in the equatorial plane of the Galactic System has been sketched. Let V_0 be the translational velocity with which the Sun rotates around the centre C. The distance from the Sun to the centre is R_0. A hydrogen cloud W is situated at a distance R from the centre; the angle at which it is observed from the Sun (or the Earth) is l' (the direction of the centre corresponds to $l' = 0$; l' is related to the galactic longitude l according to the equation $l' = l - 327.5°$). The translational velocity of the cloud is V. The component of this velocity in the direction of sight is then $VR_0 \sin l'/R$, the component of the velocity of the Sun in the direction of the cloud is $V_0 \sin l'$, and the radial component of the velocity of the cloud relative to the earth as a result of the differential rotation is thus:

$$V_{\mathrm{rad}} = \left(\frac{V}{R} - \frac{V_0}{R_0}\right) R_0 \sin l'.$$

Writing this in terms of angular velocities ω and ω_0, we get

$$V_{\mathrm{rad}} = (\omega - \omega_0)\, R_0 \sin l', \quad . \ . \ . \quad (1)$$

(A radial velocity away from the Earth is denoted positive). In the direction of the centre and in the opposite direction (anticentre) V_{rad} is zero; for all other directions however it has a finite value because $\omega \neq \omega_0$ if $R \neq R_0$. The relationship between ω and R is known approximately from statistical investigations and thus, if V_{rad} can be determined for a cloud, the distance from the centre and thus also from the Sun can be estimated.

The radial velocity can be found by studying the profile of the received 21 cm line in detail. The line is not sharp, that is to say, radiation of neighbouring wavelengths is also received. The centre of the line in general does not lie exactly at 21.1 cm, owing to the Doppler effect: the existence of a radial velocity between source and observer gives rise to an observed wavelength λ' of a spectral line which differs from the real wavelength λ in the ratio

$$\frac{\lambda'}{\lambda} = 1 + \frac{V_{\mathrm{rad}}}{c}, \quad . \ . \ . \ . \ . \ . \quad (2)$$

where c is the velocity of light. If V_{rad} is positive (object moving away), the line appears to lie at a greater wave length, and vice versa.

If we observed only one hydrogen cloud in our line of sight, and if this cloud as a whole participated only in the rotation of the galaxy and possessed no motion of its own, the spectral line would remain sharp and merely be displaced. V_{rad} could then be calculated directly from the displacement. Actually, the clouds have their own motions, with an average speed of 8 km/sec in arbitrary directions, and therefore the line is not sharp, but exhibits a broad profile. *Fig. 3a* and fig. 3*b* show profiles of the 21 cm line observed in various directions. In the direction of the centre ($l' = 0$) and the anticentre ($l' = 180°$), corresponding to the galactic longitudes $l = 327.5°$ and $l = 147.5°$, the profile is almost symmetrical; differential rotation gives rise to no radial velocity

[3]) H. C. van de Hulst, C. A. Muller and J. H. Oort, Bill. Astr. Inst. Neth. **12**, 117, 1954. For a simpler treatment, see H. C. van de Hulst, Observatory **73**, 129, 1953.

in this case. In all intermediate directions, the profile is asymmetrical. The centre of gravity of the "line" is displaced and sometimes the line shows two or three definite maxima. Correction of the asymmetrical profiles for the individual cloud motions deduced from the profiles at $l' = 0$ and $l' = 180°$, makes it possible to derive the effect of differential rotation and, at the same time the

The interstellar hydrogen is concentrated in the hatched regions, and close examination shows that these regions have the character of spiral arms. Unfortunately the measurements give less unambiguous results as soon as one is dealing with that region of the system which lies closer to the centre than the Sun; there are then two points along the line of sight where the cloud can be situated and it

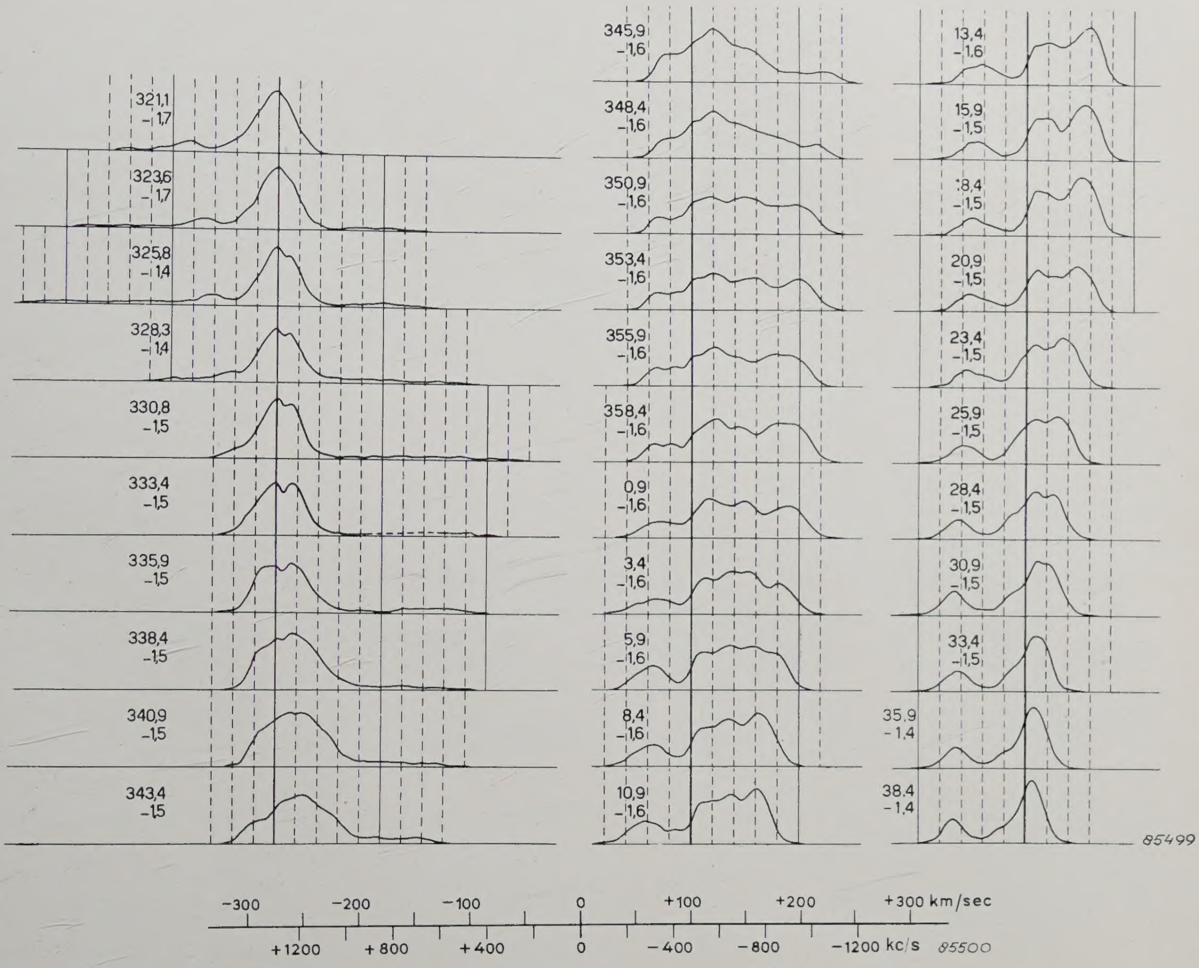

Fig. 3*a*. Line profiles of the hydrogen spectral line at 21 cm, for directions between 320° and 40° galactic longitude, i.e. near the galactic centre ($l = 327.5°$). (The negative numbers -1.4.., -1.7 represent the galactic latitude.) The deviations from the ideal profile (as found in the direction of the centre itself) depend on the Doppler effect resulting from differential rotation. The profiles can be interpreted with the help of the scale underneath the figure; this gives the relationship between relative velocity and frequency displacement.

distances of the clouds or cloud complexes from which the radiation orginates. The occurrence of more than one maximum indicates the existence of various regions of clouds situated at various distances from the Sun along the line of sight under consideration.

The distribution of clouds found in this way can be plotted on a chart. This gives a picture such as that in *fig. 4* [3]).

is very difficult to distinguish between them. Even so, indications have been found here too of a spiral structure, which is shown with dotted lines in fig. 4.

Radio-astronomical research thus lends support to the hypothesis of a spiral structure suggested by conventional astronomical work.

This is not a suitable place to go further into the many interesting results which radio astronomy has obtained in the few years of its existence. Apart

from the work on hydrogen radiation there is, for example, work proceding on continuous radiation which, amongst other things, gives us an impression of the distribution of matter generally in the Galactic System. There are also the investigations on the radiation from the Sun (chiefly originating in the corona). We shall, instead, return to our theme and discuss the basis of the Kootwijk receiver.

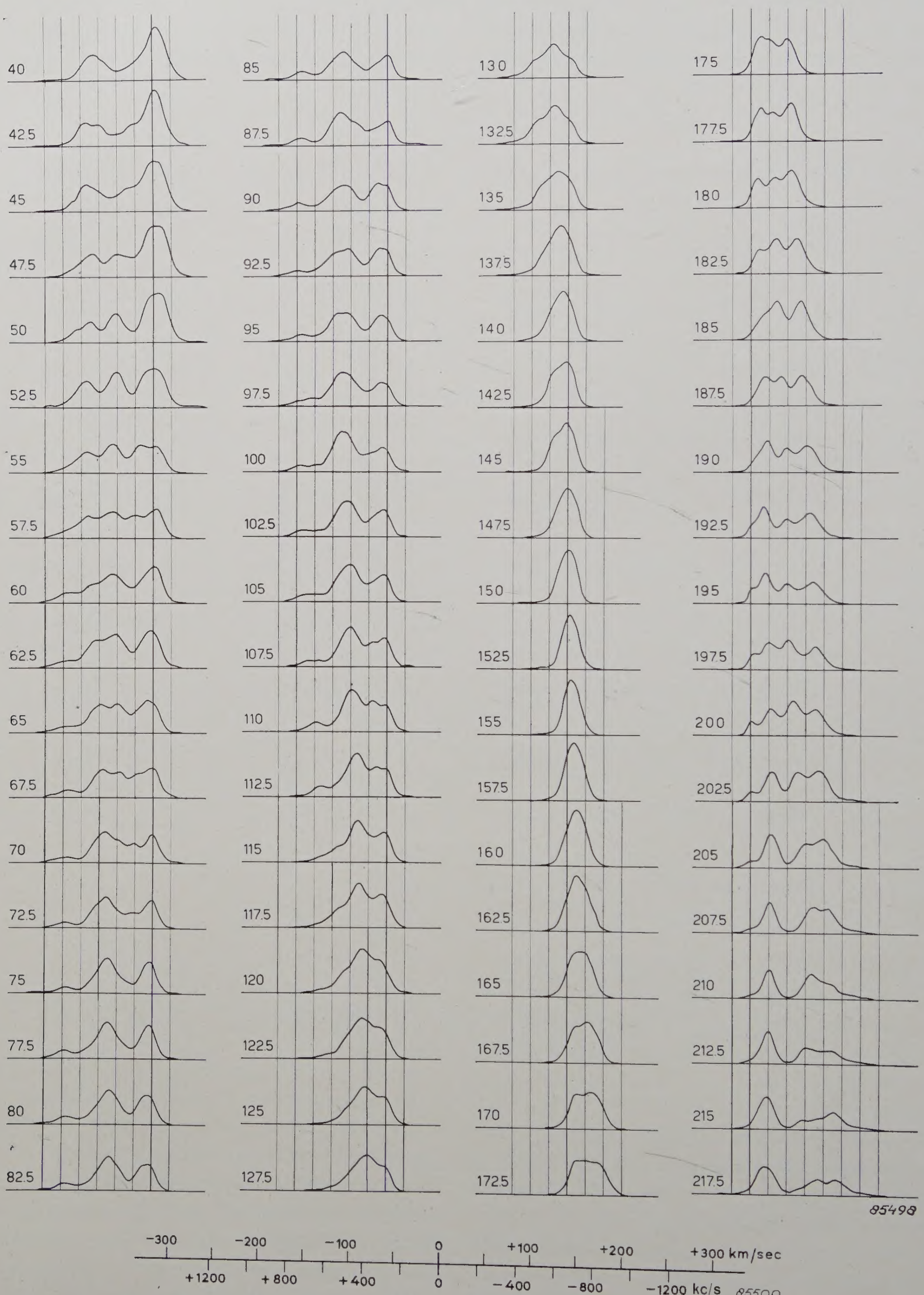

Fig. 3*b*. As for fig. 3a, for other directions in the equatorial plane of the Milky Way. The ideal profile occurs in the direction of the anticentre ($l = 147.5°$).

Properties of the radiation and design of the antenna

The aim of the investigations at Kootwijk was to measure the intensity of the hydrogen radiation as a function of position in the sky and of the frequency. In radio-astronomy, the intensity of the radiation received is usually expressed as a temperature. The energy received is imagined as coming from a closed envelope surrounding the antenna, and the temperature given is that which this envelope would have to have in order that, within the narrow bandwidth of the receiver, it should supply the same radiation power to the antenna as the latter in fact receives from space. In such a system, there is radiation equilibrium between antenna and envelope, and the

Anticentre

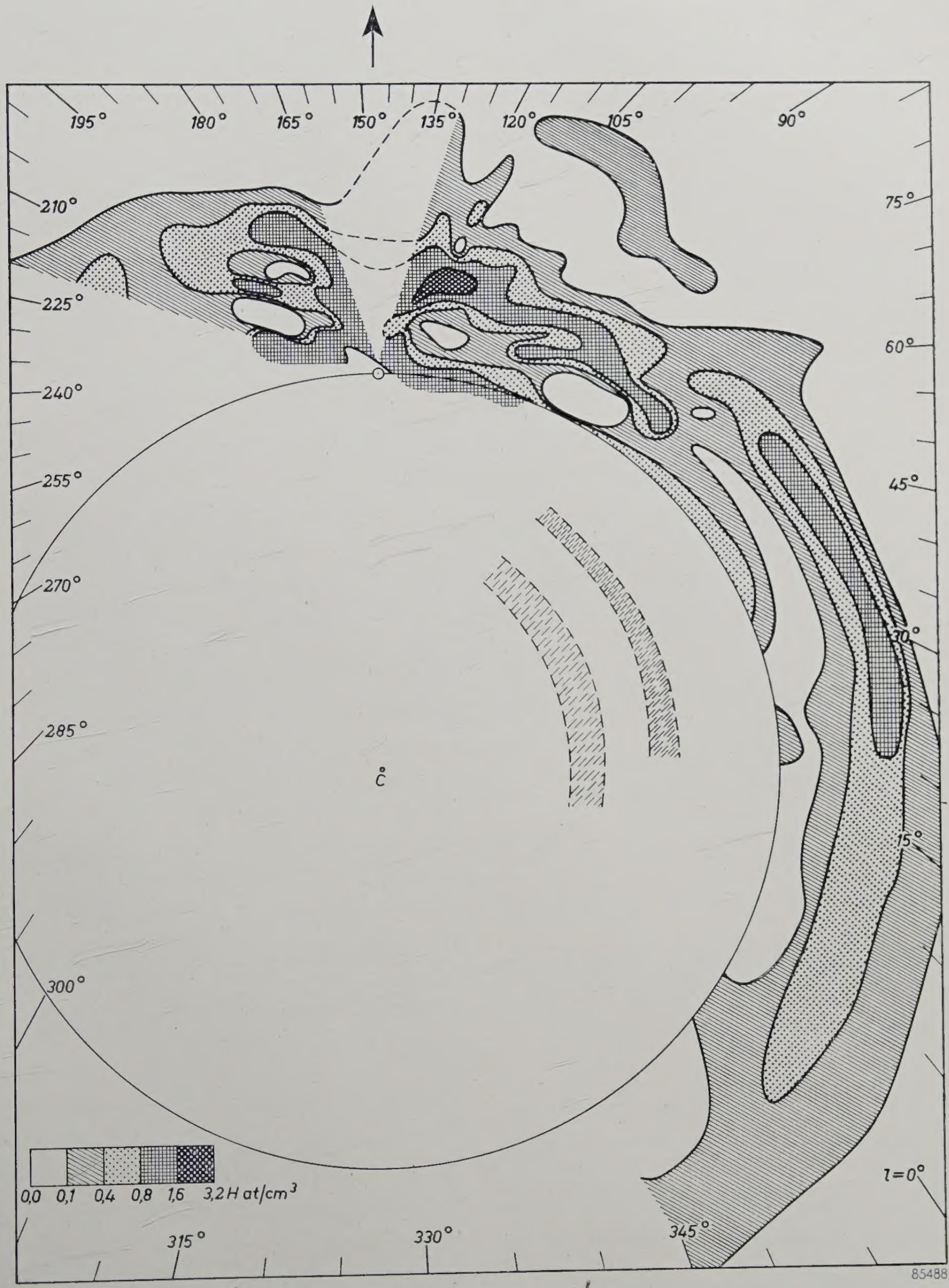

Fig. 4. Distribution of interstellar hydrogen in the equatorial plane of the Galactic System (schematic). The numbers around the edges give the galactic longitude l; the centre is indicated by C and the Sun by a small circle vertically above it. The hydrogen density is given by the shading (see code bottom left). The spiral character of the arms can be detected. Observations of spiral arms which lie closer to the centre than the Sun, are drawn with broken lines. (Taken from articles [3]) and [4])). Between $l = 130°$ and $l = 165°$ the effect of the differential rotation is so small that the measurements are not sufficiently accurate.

two temperatures are equal. It is thus usual to speak of "antenna temperature". This temperature would correspond to the kinetic gas temperature of the hydrogen clouds if they radiated as a black body (which depends partly on the thickness of the gas layer) and the antenna received radiation only from the clouds. According to various measurements, the kinetic gas temperature is about 125 °K. The highest measured "antenna temperature" is about 118°K. This corresponds, for a receiver bandwidth of 40 kc/s, to a noise power of 6×10^{-17} watts. (In the R.F. region the Rayleigh-Jeans law applies, which states that the radiation intensity of a black body is proportional to the temperature and the square of the frequency.) In the receiver to be described, temperature differences of 1 °K representing differences in power of 5×10^{-19} watts, can be measured.

The connection between "observed" intensity distributions (i.e. antenna temperature) and the "real" intensity distribution depends on the direction characteristics of the antenna used. The antenna pattern shows the sensitivity of the antenna for reception from various directions, and consists generally of a "major lobe" and a few smaller "side lobes" (*fig. 5*). The observed radiation intensity must be corrected for this antenna pattern. In practice the observed antenna temperature is taken as an average of the real temperature distribution across the half-width of the major lobe (c.f. fig. 5).

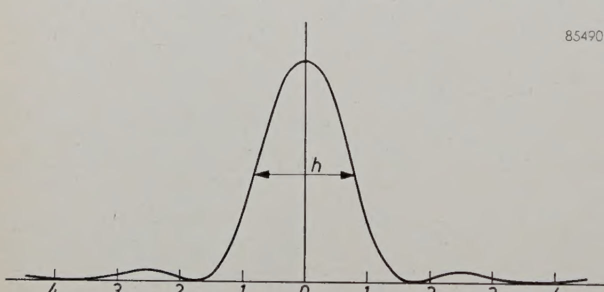

Fig. 5. Schematic directional diagram of a parabolic antenna. The ordinate represents the relative sensitivity, whilst the abscissa represents the direction relative to the axis, plotted in arbitrary units of angular measure. *h* is the half width of the "major lobe".

The finest details of the directional distribution which can be measured, are therefore fixed wholly by this half width.

The directional diagram of a parabolic reflector is almost entirely determined by diffraction phenomena. According to the theory, the half width of the major lobe (in radians) is given by:

$$\varphi \approx \lambda / D$$

in which λ is the wavelength of the radiation and D the diameter of the mirror. In the case of optical mirrors, the wavelength concerned is about $10^5 \times$ smaller than the hydrogen radio waves; and since the mirrors used in both cases differ in size by a factor no greater than 10, this means that the resolving power of a radio telescope is 10 000 to 100 000 times worse than that of an optical telescope. This is one of the disadvantages of radio astronomy.

The Kootwijk reflector (now at Dwingelo) is a paraboloid of revolution constructed of punched aluminium plate. It has a diameter of 7.5 m and a focal length of 1.70 m (*fig. 6*). Mounted at the focus is the antenna, a half-wave dipole. The observation cabin, which contains the receiver is mounted so that it turns together with the reflector in the horizontal plane. The reflector can also be moved in the vertical plane. The accuracy of adjustment is 0.1° in both directions. This is more than adequate since the half-width of the major lobe of the antenna pattern in the two main directions (along the axis of the dipole and perpendicular to it) is 2.7° and 1.9° respectively.

The frequency of the hydrogen line is accurately known from laboratory measurements [4]), viz.

$$f_0 = 1420.4056 \text{ Mc/s.}$$

The line width (half-width of the line profile) which is theoretically to be expected as a result of the thermal motion of the hydrogen atoms in a cloud, is about 5 kc/s. The Doppler effect mentioned above gives rise, in practice, to complicated profiles whose width varies between 0.05 and 1 Mc/s (cf fig. 3). Furthermore, there is in general also a weak continuous radiation from the galaxy, the temperature of which is usually much less than 10 °K, but which can rise in some directions to 60 °K. This radiation can be regarded as independent of frequency in the small frequency range under consideration. The spectrum of the galactic radiation (in this frequency range) thus looks like that shown schematically in *fig. 7*. The finest details of this profile which can be measured, are determined by the bandwidth of the receiver.

Principle of the method of measurement

In principle, the measurement of the profile of the spectral line, as sketched in fig. 7, could be done by traversing the frequency range with a receiver of very narrow bandwidth and recording the amplified power. An important factor which makes it necessary to use a more complicated measuring technique, has already been mentioned above: the intensity of the radiation received is at most only

[4]) E. G. Prodell and P. Kusch, Phys. Rev. **79**, 1009, 1950.

a few percent of the noise intensity generated in the first stages of the amplifier. In addition, the statistically fluctuating character of the received radiation makes it necessary to take each measurement over a considerable period, in order to average out the radiation intensity at a particular frequency sufficiently ($1\,^\circ\mathrm{K} \approx 0.1\%$ of the noise level in the receiver). To traverse completely the frequency range of interest takes at least an hour. However it is scarcely possible to keep the properties of the receiver constant enough for so long a time. It is therefore preferable to choose a method which avoids the effects of small changes in amplification. Since the receiver characteristics change only slowly, if a measurement is carried out within, say, 0.01 seconds, the receiver characteristics can be considered constant during this interval. Of course, the accuracy of such a short measurement is small. It can be improved, however, by taking the average result of a large number of measurements made in quick succession.

In the Kootwijk receiver, shown in *fig. 8*, each measurement consists of determining the difference between the output voltage at a frequency f_1 on the spectral line and that at an adjacent frequency f_2 (see fig. 7). Each measurement occupies a period of 1/400 sec; the receiver is switched to and fro between the two frequencies f_1 and f_2 at a switching rate of 400 c/s. To traverse the complete line profile, the two frequencies are displaced together slowly while maintaining

Fig. 6. The parabolic antenna at Kootwijk (now at Dwingelo). The reflecting surface is formed of punched aluminium sheet which is not easily visible in the photograph.

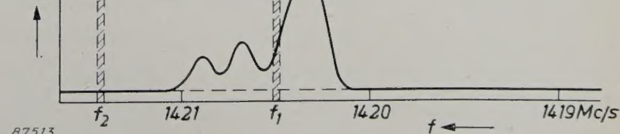

Fig. 7. Schematic representation of a line profile, i.e. of the average intensity (expressed as antenna temperature T_a) of the Galactic radiation as a function of the frequency near the 21 cm line.

a constant frequency difference. The frequency difference is so chosen ($f = 1080$ kc/s) that the comparison frequency f_2 always falls outside the spectral line. Since the (noise) signal at the frequency f_1 is always, on the average, somewhat greater than at the frequency f_2, the output voltage after rectification displays a square wave form with a frequency of 400 c/s, the amplitude of which is a measure of the difference in noise levels at the two frequencies. This method not only eliminates the effect of changes in the receiver characteristics, since each measurement lasts only a short time, and but also, since it is a difference method, suppresses the contribution of the continuous background. By measuring only the fundamental of the 400 c/s square wave voltage, with an extremely narrow frequency range on either side, the effect of the receiver noise is considerably reduced.

Fig. 8. The radio receiver as installed at Kootwijk.

Since during each half period of the measurement the receiver is tuned to a frequency outside the spectral line, the available information time is only half used, and the accuracy of the measurement is thus reduced by a factor 2. To improve on the usage of the available time, two comparison frequencies are used in practice. A more detailed explanation and some of the circuit details are given in the second part of this article.

The amplitude of the 400 c/s output signal is recorded on a strip-chart recorder in which the chart movement is correlated with the frequency. *Fig. 9* shown a typical recording obtained on the Kootwijk equipment.

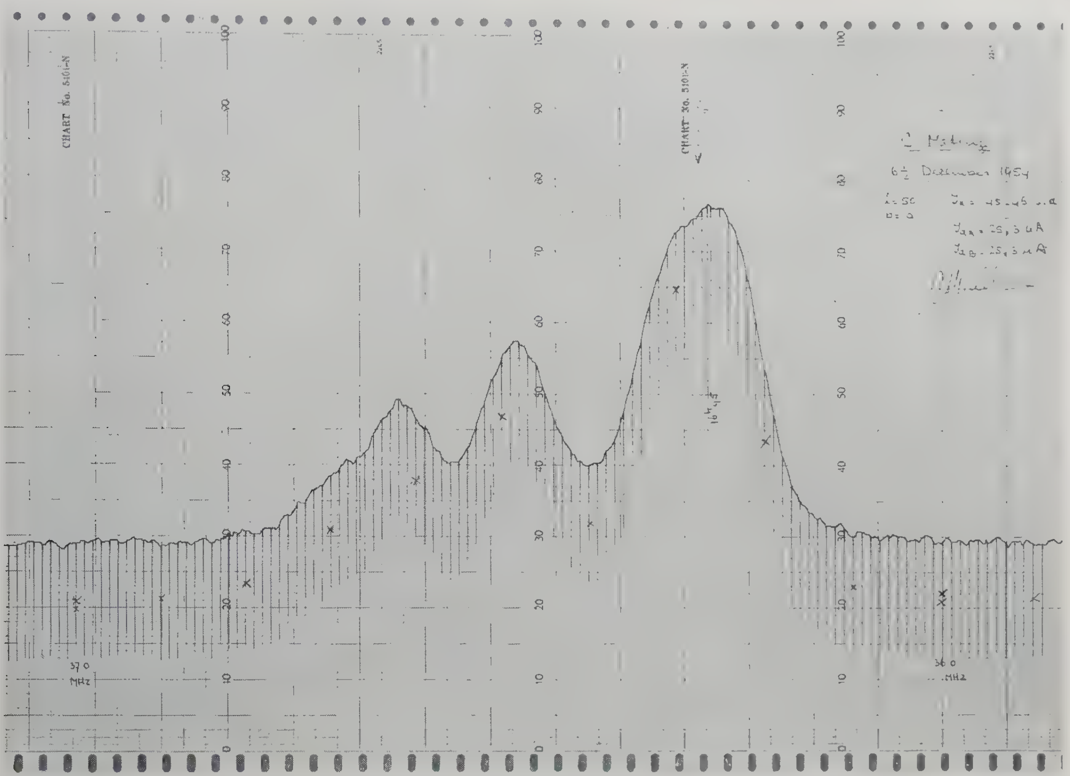

Fig. 8. Example of a record. The vertical lines at 10 kc/s intervals mark the frequency.

Summary. Radio astronomy has grown in a very short time to be a powerful tool among astronomical techniques. This article deals briefly with the structure of the Galactic System and the role played by interstellar hydrogen as source of radio waves of sharply defined frequency; this is followed by a short description of the principles of operation of a receiver used to detect these waves and measure their intensity set up at Kootwijk (and now removed to Dwingelo) by the Netherlands Foundation for Radio Astronomy. The hydrogen radiation from space is picked up by a dipole antenna in the focus of a parabolic reflector 7.5 m in diameter. The radiation has the character of noise and is very weak, many times weaker than the noise level in the receiver itself; this necessitates special methods of measurement in order to achieve a reasonable accuracy. If we express the received noise power as a temperature defined by means of the Rayleigh-Jeans law, the maximum received intensity corresponds to a temperature of about 110°K. The accuracy of measurement reached is 1°K, i.e. 0.1% of the receiver noise.

A TRANSISTOR HEARING-AID

621.395.92:621.375.4

The most obvious advantages of the transistor over the thermionic valve are that the dimensions and the power consumption are considerably smaller. It requires, moreover, only one source of energy of very low voltage (a few volts only); and in comparison with electronic tubes there is not only the advantage (where dry batteries form the source) of lower power consumption but also the fact that the energy (the number of watt hours) is supplied more cheaply from a battery of a few volts than from a battery of some tens of volts. Where hearing-aids are concerned, small dimensions and low battery costs are of course two of the most important characteristics. It is hardly surprising therefore, that one of the first applications of Philips transistors has been in hearing-aids.

The first Philips transistor hearing-aid (type KL 5400) has meanwhile been superseded by a second (type KL 5500, *fig. 1*) in which the specific properties of transistors are even better employed. This apparatus, now on the market for about a year, weighs only half as much as the hearing-aid using

valves, earlier described in this Review [1]), whilst running costs (batteries) have been reduced to $^1/_{10}$. Another appreciable advantage of the new apparatus is that it provides a greater acoustical gain and produces a higher sound level than the valve hearing-aid; a wider circle of those who are hard of hearing may therefore profit from it.

The hearing-aid KL 5500 consists of an electromagnetic microphone, an amplifier with four transistors, a telephone inset and a battery supply of one, two or three 1.2 V cells. The nominal acoustical gain is 58 dB with one cell and 69 dB with three cells, corresponding to a power output of the amplifier of 1 and 10 mVA respectively.

The transistors (three of type OC 70 and one, in the output stage, of type OC 71) are resistance-coupled and connected according to the common emitter type of ciccuit [2]). The output transistor supplies its power directly to the telephone without the intermediary of an output transformer as required with a tube. The omission of an output transformer is possible because of the inherently low output impedance of the transistor circuit in question [2]). As a rule all impedances of transistor circuits are lower than those of circuits with electronic tubes. A consequence of this is the great advantage of transistors that undesired couplings due to stray capacitances are negligible, so that a compact construction can be used. Since transistors, unlike valves, exhibit no microphony, no special provisions are required to counteract this effect; this also contributes towards a compact construction.

For a hearing-aid, a good limiter in the output stage is essential; this is to prevent the threshold of pain of the wearer being exceeded in case of shouting or other loud noises; patients with reduced span are particularly susceptible to this trouble. In the hearing-aid KL 5500 limiting is obtained by selecting appropriate values for the current and the impedance in the collector circuit of the OC 71. For further particulars we refer the reader to a more detailed article to appear in this Review.

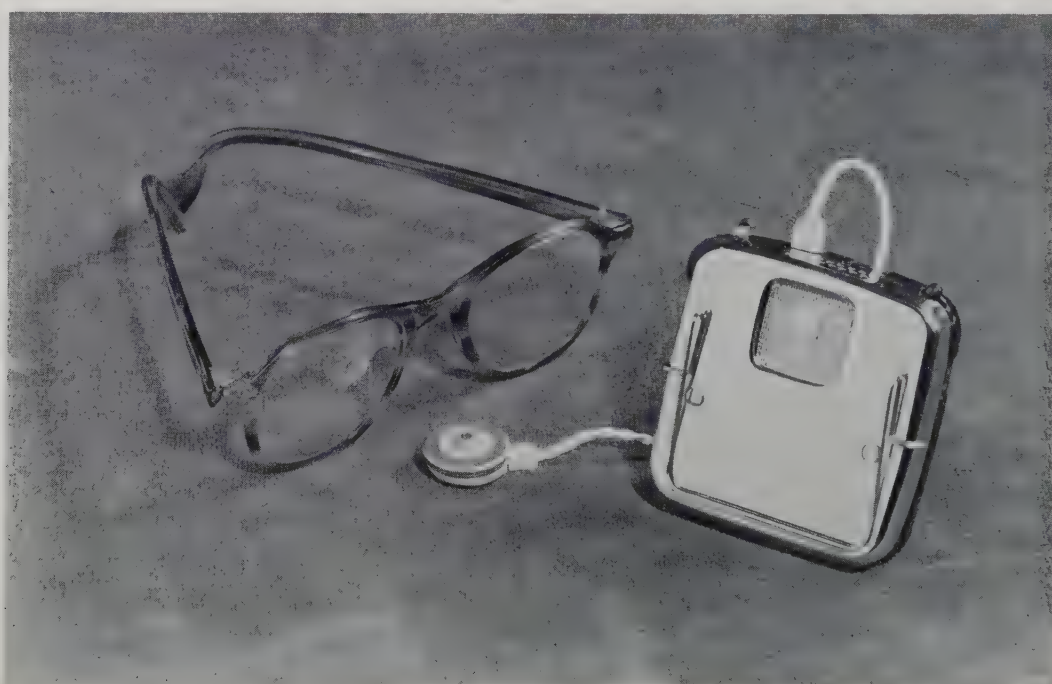

Fig. 1. The hearing-aid KL 5500 which works on transistors.

In order to provide a sound that is agreeable to the wearer, one of the essential requirements is a smooth and adjustable response curve. An uneven response curve is mainly due to resonances of microphone and telephone. The response curves of both have been improved by providing acoustical cavities with damping. The sensitivity is impaired by this, but a higher sound level than that of a valve hearing-aid can nevertheless be readily obtained. If necessary, the power output can be increased to the relatively high value of 10 mVA, by using three cells (3.6 V).

As far as the adjustment of the response curve is concerned, the wearer has first of all the choice of

[1]) P. Blom, An electronic hearing-aid, Philips tech. Rev. **15**, 37-48, 1953/54.

[2]) See, for example, Philips tech. Rev. **17**, 242, 1955/56 (No.9).

three telephones with different characteristics; moreover, the response curve of the amplifier can be adjusted by means of a three-position selector knob. A choice of nine response curves is thus available. As in all fields of communications, the signal-to-noise ratio is an important quantity. The first transistors were, as regards noise, somewhat inferior to electronic tubes, but the modern types have been so much improved that at present a transistor apparatus can be made having less noise than a good valve hearing-aid. The equivalent input signal for noise in a valve hearing-aid is approximately 25 dB above zero level (10^{-16} W/cm^2), whereas in the hearing-aid KL 5500 this quantity is less than 20 dB.

Not only the electronic noise but also the case noise has to be considered. This latter is caused by the fact that the ear of the patient lies, as it were, on his clothes. This noise shows a fairly uniform distribution throughout the spectrum, and the level of interference thus caused corresponds roughly to the level of normal speech at a distance of 1 metre. In the KL 5500 this noise has been considerably reduced by acoustical insulation of the microphone cavity from the rest of the apparatus.

P. BLOM.

BEAM TRANSMITTERS WITH DOUBLE FREQUENCY MODULATION

by C. DUCOT *).

621.396.43:621.376.32:621.396.5

The development of the very-high-frequency techniques (wavelengths of a few centimetres) has made it possible to establish carrier-telephony communication by means of accurately directed radio beams. Equipment developed for this purpose will be described in a number of articles in this Review. It will be shown that the application of double frequency modulation results in a favourable signal-to-noise ratio, and also provides some practical advantages important to the telephone service.

Beam transmitters operating on wavelengths between a few centimetres and a few metres are being increasingly used for the transmission of telephone calls and TV-signals. Two methods of modulation may be adopted for this, viz. frequency modulation [1]) and pulse modulation [2]). For transmitting a large number of telephone channels or one or more TV-signals, only frequency modulation can at present be employed. In this system a group of speech channels is transmitted as a single signal, in the same way as in carrier telephony via cables. A beam-transmitter link using frequency modulation may therefore take the place of a normal cable circuit without any complications. At large distances between transmitter and receiver the input signal of the receiver is necessarily attenuated, so that the ratio between the desired signal and the inevitable noise deteriorates. This noise, which originates mainly in resistors and electronic tubes, will be referred to here as *fluctuation noise*. Apart from this, in carrier telephony, other interference voltages are added to every channel, owing to the fact that the transmission of a signal always causes some distortion.

This distortion is mainly due to the fact that the modulation characteristic of the frequency modulator in the transmitter as well as the demodulation characteristic of the discriminator in the receiver are never perfectly straight. Furthermore, a certain distortion arises because the phase-characteristic of the networks through which the signal passes in transmitter and receiver, are not perfectly linear. Finally, modulation distortion may arise from the signal being transmitted via a long cable (e.g. an antenna feeder) at least one extremity of which does not have a reflection-free termination, i.e. a resistance equal to the characteristic impedance of the line. The distortion due to the two latter causes is often termed phase distortion. When the signal is transmitted through a wave guide, even a perfectly reflection-free termination cannot entirely prevent phase distortion. Finally it should be mentioned that modulation distortion can also arise when the electromagnetic waves from the transmitter reach the receiver along different paths.

In the case of a large number of channels, which carry widely varying loads in the course of normal telephonic traffic, interfering alternating voltages are manifested as noise in every channel. In accordance with the usual terminology this noise will be referred to as *intermodulation noise* [3]).

In this article we shall deal with a method which makes it possible in many cases to reduce the noise

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[1]) Cf. A. van Weel, An experimental transmitter for ultra-short-wave radio telephony with frequency modulation, Philips tech. Rev. **8**, 121-128, 1946.

[2]) Cf. C. J. H. A. Staal, An installation for multiplex-pulse modulation, Philips tech. Rev. **11**, 133-144, 1949.

[3]) See: J. te Winkel, Feedback amplifiers for carrier telephone systems, Philips tech. Rev. **16**, 287-296, 1954/55.

in a radio link, thus enabling either greater distances to be bridged or more channels to be transmitted.

The method consists of so-called double frequency modulation, in the sense that first an auxiliary carrier wave is modulated with the signal, after which the main carrier wave, generated by an oscillating transmitting valve, is modulated with the modulated auxiliary carrier.

By way of introduction, we shall first consider the requirements to be met by a radio link for carrier telephony. After discussing the method of double modulation we shall finally describe an experimental installation operating on this principle.

Criteria for a radio link

During a congress held in Florence in October 1951, the C.C.I.F. (Comité Consultatif International de Téléphonie) recommended that beam-transmission links for carrier telephony should conform, whenever possible, to the standards applying to international cable telephone circuits. These standards are defined with the aid of a "standard circuit". This consists of a transmission path of total length 2500 km, split up at nine places where demodulation into the supergroups (each containing 5 groups of 12 channels) and renewed modulation takes place.

Owing to attenuation and the interposition of line amplifiers (repeaters) the signal power varies considerably along a telephone line (e.g. the power is low at the input side of a line amplifier and high at the output side). At certain places the power is equal to that at the place of origin [4]); these places are then said to be at *relative zero level*. The strength of a signal is often expressed in decibels relative to a signal having a power of 1 mW at a point at relative zero level, and denoted by dBm_0 (m standing for mW and 0 for zero level).

One of the criteria for the above-mentioned standard circuit is that the level of the total noise in each of the channels, in so far as it originates from that particular line section, and measured psophometrically [5]) at a place with relative zero level, must not exceed a value of 7500 pW for more than 1% of the time. In this connection it must be observed that with radio links, it is necessary to take account of a special condition not found with cable circuits, viz. the occurrence of fading. When frequency modulation is used, the effect of fading is not that the damping of the circuit increases, since the magnitude of the alternating voltages obtained in the receiver after frequency demodulation is determined by the frequency swing, and this swing is not affected by fading. Fading, on the other hand does cause an increased fluctuation noise, thus reducing the signal-to-noise ratio.

When assessing the fluctuation noise in circuits incorporating one of more radio links, therefore, the incidence of fading should be reckoned with. The effect of this, however, is the same for the two systems which we shall compare here (viz. single and double modulation).

When a telephone circuit includes a radio link, there is a fixed relation between the signal level at any point of the circuit and the frequency swing of the beam transmitter. A quantity characteristic of the link is, therefore, that particular value of the frequency swing which produces a signal strength of 1 mW in the points at relative zero level. The r.m.s.-value of this frequency swing will be denoted d.

There is considerable freedom in the choice of d, since by varying the amplification before the modulator stage in the transmitter and that after the demodulator in the receiver in opposite directions, the damping of the whole circuit can be kept constant at a variable value of the frequency swing. By increasing the frequency swing, the ratio of signal to fluctuation noise can be improved. This involves, however, a greater distortion, which, as we explained above, manifests itself in an increase in the intermodulation noise. A certain value of d may be found where the sum of these two noise components is a minimum, and the frequency swing should preferably be adjusted to this optimum value. It would, therefore, be illogical to compare the properties of two systems solely on the basis of either the fluctuation noise or the intermodulation noise. The properties of a link can, however, be expressed unambiguously by the noise occuring at the optimum frequency swing.

As stated, intermodulation noise is caused by the fact that the signals are distorted in transmission. A measure of this noise, therefore, is the relative strength of the higher harmonics when a purely sinusoidal signal is transmitted instead of the composite signal from all telephone channels. In many cases it will be sufficient to consider only the second harmonic [6]). In that event the total noise

[4]) See e.g. fig. 1 in J. de Jong, Maintenance measurements on carrier telephony equipment, Philips tech. Rev. **8**, 249-256, 1946.

[5]) The psophometrically measured power (which takes into account the frequency-dependent sensitivity of the human ear) can be found with an accuracy sufficient for our pupose by measuring the power in the channel in question (frequency band 4 kc/s) with a frequency-independent instrument, and reducing the result by 3 dB.

[6]) Only in the rare event that the channels are at frequencies so high that the whole frequency band of all channels is contained in one octave will the part played by the second harmonics and the corresponding combination frequencies be of little importance.

power W_b, measured psophometrically in one channel, at a point at relative zero level, can be expressed by the formula:

$$W_b = \frac{A}{d^2} + Bd^2, \quad . \; . \; . \; . \; . \; . \quad (1)$$

where A and B are constants [7]).

The fluctuation-noise level in the frequency band of one channel may be expressed by a specific frequency swing. The mean square value of this frequency swing is represented by the constant A. This quantity is not dependent upon the earlier introduced quantity d (which determines the effective signal), but does depend upon the power of the transmitter, the directional properties of the antennae, the conditions of propagation of the electromagnetic waves, and furthermore upon the noise factor [8]) of the receiver, the number of transmission links and upon the central frequency of the frequency band available for the channel in question.

The intermodulation noise, on the other hand, does depend upon d; its contribution to the total noise power in each channel is in fact proportional to d^4. The proportionality constant B is dependent upon the various factors causing the distortion, upon the value of the total signal of all channels, and (like A) upon the central frequency of the frequency band available for the channel in question.

The total noise level in milliwatts in a point at relative zero level is then obtained by dividing the sum of A and Bd^4 by d^2, which corresponds to formula (1).

This expression is a minimum for a value of d given by:

$$d_0 = \sqrt[4]{\frac{A}{B}}, \quad . \; . \; . \; . \; . \; . \; . \; . \quad (2)$$

and the minimum value of the noise power amounts to:

$$W_{b\,min} = 2\sqrt{AB}, \quad . \; . \; . \; . \; . \quad (3)$$

This quantity determines the properties of the transmission link. (Communication is better according as the above quantity is smaller).

It should be observed that, where minimum noise is concerned, i.e. where $d = d_0$, the contributions of fluctuation noise and of intermodulation noise towards the total noise are the same. A great difference between these quantities is, therefore, an indication that the frequency swing d has not been appropriately chosen.

Double frequency modulation

One of the major difficulties in the development of a radio link is maintaining the linear relation between the signal applied to the transmitter and the signal obtained in the receiver after demodulation. Any deviation from this linearity gives rise to distortion and thus, as we explained above, to intermodulation noise. A first requirement, therefore, is that modulation in the transmitter is effected linearly, i.e. that the difference between the momentary frequency and the central frequency is exactly proportional to the voltage applied to the modulator. Circuits to realize this, however, can only operate at relatively low frequencies (up to approx. 150 Mc/s) with present facilities. The far higher frequencies that are, as a rule, required for the signal to be transmitted, may then be obtained by a system of frequency transformation. A block diagram of a transmitter operating on this principle is shown in *fig. 1a*. The signal to be transmitted is indicated here by S (frequency f_s). The oscillator O_1 (frequency f_1) is modulated by the frequency modulator FM. The modulated voltage is applied to the mixing tube M, in which, with the aid of the oscillator O_2 (frequency f_2) a voltage with the far higher central frequency f_3 is formed ($f_3 = f_2 + f_1$) this voltage, after being amplified in A_H, is applied to the transmitting antenna A_T. The output tube of the transmitter must be an amplifying tube, and the tubes now available for the high frequencies involved (e.g. 4000 Mc/s) can supply only a low power (1 W or less). This limited output power has an unfavourable effect on the ratio of signal to fluctuation noise.

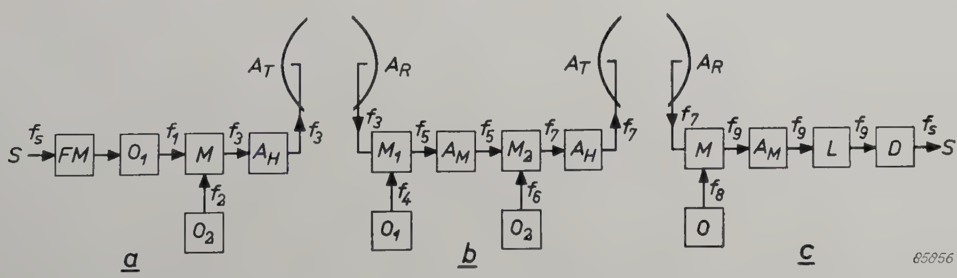

Fig. 1. Block diagrams of a transmitter (a), a relay station (b), and a terminal station (c), using an amplifying tube in transmitter and relay station. The significance of the letters is given in the text.

7) See also: A. T. Starr and T. H. Walker, Microwave radio links, Proc. Instn. El. Engrs. **99**, 241-255, 1952 (partIII).

8) For the significance of this term, see e.g.: G. Diemer and K. S. Knol, The noise of electronic valves at very high frequencies, II, The triode, Philips tech. Rev. **14**, 236-244, 1952/53.

When an oscillator is used as transmitting tube, a far higher output can be obtained, viz. of the order of 10 W. The signal to be transmitted is in this case directly applied to the transmitting tube O (*fig. 2a*). In this tube, frequency modulation with a very good linearity should be effected. It has been mentioned earlier, however, that this requirement cannot be satisfied at high frequencies. (Strong negative feedback could be applied, but in that case it is very difficult to obtain and maintain a stable adjustment of the transmitter). The block-diagrams *b* in figs. 1 and 2 represent relay stations, which relay the signals if more than one radio link is used. In the two cases (1*b* and 2*b*) totally different circuits have to be applied. If an amplifying transmitting tube is used (fig. 1*b*), an I.F. voltage (frequency f_5) can be obtained by combining the H.F. signal received in antenna A_r with the output of oscillator O_1 (frequency f_4) in the mixing stage M_1. This I.F. voltage, after being amplified in the I.F. amplifier A_m, is applied to a second mixing stage M_2. In this stage, by means of the oscillator O_2 (frequency f_6), an H.F. signal (frequency f_7) is again obtained which, after being amplified by amplifier A_H, is applied to the transmitting antenna A_T. As a rule f_7 is made different from f_3 so as to prevent many undesirable coupling effects between input and output. In relay stations it is therefore not always necessary to return to the A.F. signal by means of a frequency demodulator (as *is* obviously necessary when either telephones or a telephone circuit are connected to the relay station). If, however, an oscillator is used as the transmitting tube (fig. 2*b*), demodulation then has to be effected in every relay station. This is done by applying the I.F. amplified signal, via a limiter L, to the frequency demodulator D. The A.F. signal thus obtained is then again applied, via the amplifier A_s, to the oscillating transmitting tube O_2. Since the process of demodulation and modulation is always accompanied by some distortion, this can be regarded as a serious drawback of using an oscillator as the transmitting tube.

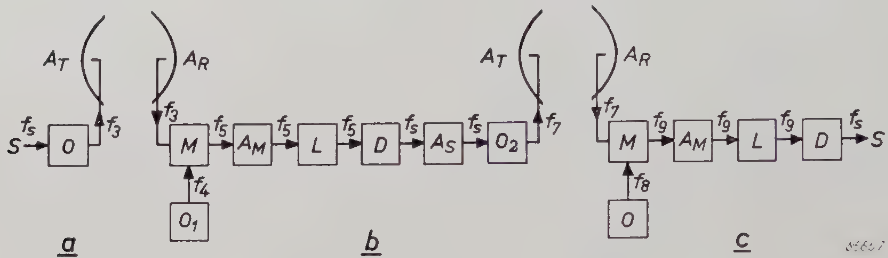

Fig. 2. Block diagrams of a transmitter (*a*), a relay station (*b*), and a terminal station (*c*), using an oscillating transmitting tube and a system of single modulation. The significance of the letters is given in the text.

The block diagrams of figs. 1*c* and 2*c* represent terminal stations. It will be clear that these may be identical for both systems. In the mixing tube M, the I.F. signal with frequency f_9 is obtained from the antenna signal with frequency f_7 and the oscillator signal with frequency f_8. This I.F. signal is applied, via the I.F. amplifier A_M and the limiter L, to the frequency demodulator D, which produces the A.F. signal S.

In *figs. 3* and *4* the various frequencies mentioned in connection with figs. 1 and 2 are represented on a frequency scale. The sections *a*, *b* and *c* are correspondingly related to the transmitter, the relay station and the terminal station.

Owing to the high power available when an oscillating transmitting tube is used, a radio link

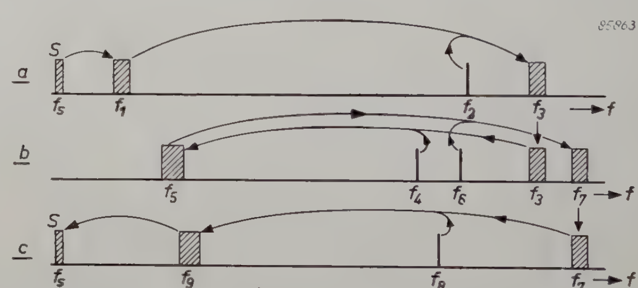

Fig. 3. Schematic representation of the frequencies of the signal voltages in a transmitter (*a*), a relay station (*b*), and a terminal station (*c*), for a circuit according to fig. 1.

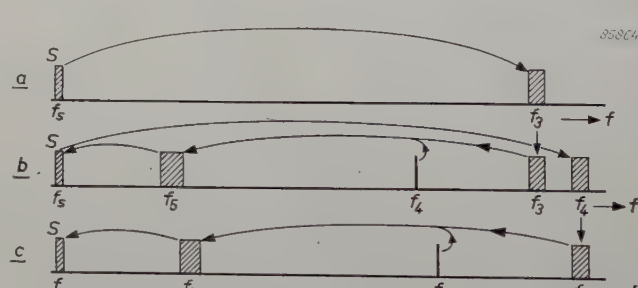

Fig. 4. Schematic representation of the frequencies of the signal voltages in a transmitter (*a*), a relay station (*b*) and a terminal station (*c*), for a circuit according to fig. 2.

operating on this principle is less likely to be affected by interference due to fading. A certain degree of fading may affect transmission by a system of oscillating transmitting tubes (fig. 2) only as a reduction of the signal-to-noise ratio, whereas under similar conditions, communication via a system with amplifying transmitting tubes (fig. 1) may be completely interrupted.

A system of modulation using an oscillator as transmitting tube and yet permitting modulation at a far lower frequency is the system of double modulation, invented by Armstrong. The application of this system to beam-transmitters was, to our knowledge, first suggested by L. E. Thompson [9]) and put into practice by the Radio Corporation of America (RCA) for an experimental communication link between New York and Philadelphia [10]). This system was subsequently applied by Western Union Telegraph for a communication chain linking New York, Washington and Pittsburgh [11]). *Fig. 5* shows the block diagram of a transmitter, a relay station and a terminal station operating on this principle, whilst *fig. 6* again represents the frequencies of the currents and voltages occuring at various points of the circuit. In the transmitter, an auxiliary carrier-wave with frequency f_1 is modulated with the signal S to be transmitted by means of the frequency modulator *FM*. The auxiliary carrier-wave is generated in the oscillator O_1. After being modulated it is applied, via the amplifier A_{SP}, to the oscillating transmitting tube O_2 (frequency f_3), in which f_3 is frequency-modulated demodulator D and applied, via amplifier A_{SP}, to the oscillating transmitting tube O_2. Demodulation of the auxiliary carrier, therefore, is not necessary in a relay station. The first stage of a terminal station is likewise a normal FM-receiver (cf. fig. 2*c*). The auxiliary carrier obtained from the frequency

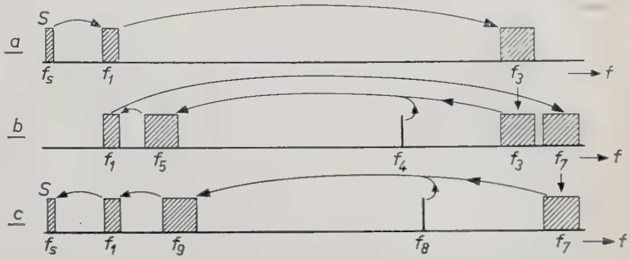

Fig. 6. Schematic representation of the frequencies of the signal voltages in a transmitter (*a*), a relay station (*b*), and a terminal station (*c*), for a circuit according to fig. 5.

demodulator D_1 is applied, via the amplifier A_{SP} and the limiter L_2, to a second frequency demodulator D_2, which finally produces the A.F. signal S.

Owing to the fact that the frequency f_1 of the auxiliary carrier must always be several times higher than the highest frequency of the A.F. signal S to be transmitted, the ratio of signal to fluctuation noise is unavoidably impaired when double modulation is applied. In the previous section, however, it was explained that it is not fair to compare two modulation systems solely with respect to the fluctuation noise. The application of double modulation unquestionably offers certain substantial advantages, viz:

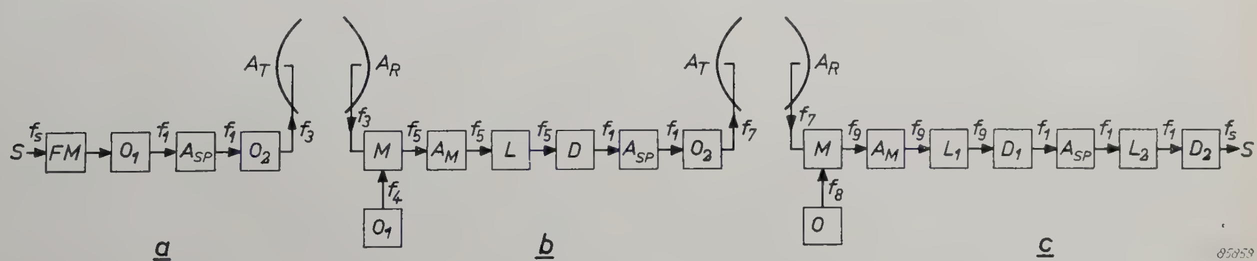

Fig. 5. Block diagram of a transmitter (*a*), a relay station (*b*), and a terminal station (*c*), using an oscillating transmitting tube and a system of double modulation.

with the modulated auxiliary carrier frequency. In a relay-station, the first stage of which is again a normal FM-receiver (cf. fig. 2*b*) the auxiliary carrier frequency is obtained by means of the frequency

1) Good linearity is required only in the first modulator and in the final frequency demodulator, both operating at a relatively low frequency. The modulation distortion affecting the carrier frequency or I.F. signals produces after demodulation only higher harmonics of the auxiliary carrier frequency. Since the carrier has to pass a filter before being applied to the second demodulator this distortion will affect the intermodulation noise to a very small degree only. For this reason no

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[9]) L. E. Thompson, A microwave relay system, Proc. Inst. Rad. Engrs. **34**, 936, 1946.

[10]) G. G. Gerlach, Microwave relay communication system, R.C.A. Review **7**, 576-600, 1946.

[11]) J. J. Lenehan, A radio relay system employing a 4000 Mc/s 3-cavity klystron, Western Union tech. Rev. **6**, 111-116, 1952.

stringent demands as regards linearity are made on the phase characteristics of H.F. and I.F. circuits in transmitters, relay stations and receivers. The same applies to the linearity of the modulation characteristics of the oscillating transmitting tubes of transmitter and relay stations, and to the characteristics of the frequency demodulator in the relay stations and of the first frequency demodulator in a terminal station. Also the phase distortion arising from the H.F. signals being transmitted via long antenna feeders will ultimately not play an important role. A relay station in which the auxiliary carrier is not demodulated, will therefore contribute very little to the intermodulation noise, a property which is not shared by either of the other systems discussed here.

2) A consequence of the points mentioned under 1) is that the frequency swing effecting modulation of the carrier with the auxiliary carrier is not greatly restricted by considerations of linearity. A wide frequency swing (e.g. 5 Mc/s or even wider) is thus permissible, which again favourably influences the ratio of the signal voltage to the fluctuation noise, whereby the adverse influence of the high frequency of the auxiliary carrier on this ratio is largely compensated.

3) With double modulation the requirements put on the limiters are less exacting than those with single modulation [7]).

4) Modulation and demodulation of the auxiliary carrier is necessary only in those places where this is required by the telephone service. The distortion, and hence the intermodulation noise is thus virtually solely determined by the number of relay stations where this *has* to take place, and not by the total number of relay stations (the latter often being five times more numerous than the former).

5) Double modulation offers the possibility of transmitting, apart from the "normal" speech channels, also a number of signals with single modulation, e.g. a service channel or a signal for measuring and inspection purposes, which require less exacting demands as regards freedom from noise. (The advantages of double modulation obviously do not apply to these channels.)

We should mention here that the application of double modulation was particularly attractive to us, as we have at our disposal an oscillating transmitting tube of extremely simple and sturdy construction, viz. the multi-reflex klystron, designed by F. Coeterier. Modulation with the wide frequency sweep in question presents no difficulties with this tube, because its "electronic bandwidth" amounts to about 30 Mc/s. A multi-reflex klystron was described some years ago in this Review; in this issue an article appears in which later versions of this type of tube are discussed [12]).

Comparison of the signal-to-noise ratios for systems with single and with double modulation

We shall now make a quantitative comparison of the signal-to-noise ratios for beam-transmission systems with single and with double modulation. It will be clear from the foregoing that in the former system the transmitting tube is assumed to be of the amplifying type (fig. 1).

For making the comparison, let us suppose that with either of these beam-transmission systems a standard circuit is formed similar to that referred to in the C.C.I.F. specifications regarding the permissible noise. This circuit has, therefore, a length of 2500 km, and consists, as will be assumed here, entirely of beam-transmission links. In compliance with C.C.I.F. specifications, we shall assume that in nine of the relay stations, demodulation into the super-groups and renewed modulation takes place.

In order to study the two systems as far as possible under equivalent conditions, all characteristic quantities not depending upon the choice of system, will be assumed to be equal in either case. These quantities are: number and length of the links, the carrier-wave frequency, the gain of the antenna system, the conditions of propagation of the electromagnetic waves between transmitting and receiving antenna, and the noise factor of the receivers. It was mentioned earlier that in view of the available tubes, the power of the transmitters of the two systems cannot be the same. Also the frequency swing will be different for either case. As regards the latter, we shall assume that in the system according to fig. 1 this is adjusted to the optimum value with a view to the noise, whilst in the system according to fig. 5 the same applies to the frequency swing upon the auxiliary carrier. In the latter case the frequency swing on the transmitted carrier will be adjusted to the maximum value permissible for the transmitting tube in question.

When a sinusoidal signal is transmitted instead of a signal composed of telephone channels, it will likewise undergo a certain distortion, or, in other words, higher harmonics will be formed. We shall now introduce the symbol μ to designate that

[12]) F. Coeterier, The multireflection tube, a new oscillator for very short waves, Philips tech. Rev. 8, 257-266, 1946. The multi-reflex klystron as a transmitting valve in beam transmitters, Philips tech. Rev, this issue, p. 328.

frequency at which, for single modulation, a specific ratio of the second harmonic to the fundamental occurs (we may take this ratio as, say, 10^{-6}, i.e. -60 dB). With double modulation the frequency swing on the auxiliary carrier which causes the same distortion will be denoted μ'. We shall further introduce the following notation: P is the output power of the transmitter when an amplifying transmitting tube is used (fig. 1), and P' is the output power radiated by an oscillating transmitting tube (fig. 5). f_1 represents the central frequency of the auxiliary carrier and Δ the frequency swing with which the auxiliary carrier is modulated upon the transmitted carrier.

It can now be demonstrated [13]) that the ratio of the total noise power of a system according to fig. 1 to that of a system according to fig. 5 can be expressed by:

$$\eta = \frac{1}{\sqrt{2}} \frac{\mu'}{\mu} \frac{\Delta}{F} \sqrt{\frac{P'}{P}} . \quad . \quad . \quad . \quad . \quad . \quad (4)$$

If the value of η turns out to be greater than 1, a system with double modulation is clearly preferable, but even if the value of η were unity or slightly less, double modulation would still be preferable as a rule, in view of some of the advantages enumerated above, which find no expression in formula (4).

If, for instance, 60 telephone channels are to be transmitted then the following values are valid for the quantities in (4):

$$\Delta = 5 \quad \text{Mc/s},$$
$$F = 1.5 \ \text{Mc/s}.$$

If, further, $P'/P = 10$, which is about right for the tubes available at present, it follows from (4) that

$$\eta = 7.5 \frac{\mu'}{\mu} . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

On the assumption that the intermediate frequency of a system with single modulation is 100 Mc/s, we may put the value of μ at about 0.35 Mc/s. From (5) it follows now that if η is to be greater than 1, μ' must be greater than 46.5 kc/s. This can easily be realized [14]) with a frequency demodulator tuned to the frequency of the auxiliary carrier (e.g. 1.5 Mc/s).

[13]) See: C. Ducot, Procédé technique pour l'amélioration des performances des faisceaux hertziens en téléphonie, Onde Electrique **35**, 41-54, Jan. 1955 (No. 334).

[14]) It would be possible to shift the frequency of the auxiliary carrier to, say 13 Mc/s by frequency transformation in the receiver, as was effected by Thompson, thus easily satisfying the above requirement for μ', but in most cases this provision is not necessary.

Experimental installation with double modulation

In order to test the advantages of double frequency modulation for beam transmitters employing the multi-reflex klystron transmitting tube, the Laboratoires d'Electronique at de Physique appliquées, Paris, have developed an experimental installation, consisting of a transmitter and a receiver.

Fig. 7. The experimental transmitter for double frequency modulation. To the left is the rack, to the right, mounted on a tripod stand, the reflector assembly, containing the output stages of the transmitter.

Each is made up of two parts, one built into a paraboloid reflector assembly, and the other built into a rack and connected via cables to the former.

Fig. 7 shows the complete transmitter, which is represented by the block diagram of *fig. 8*. The frequency of the auxiliary carrier is 2 Mc/s. Since linear modulation with the required frequency swing will present difficulties at such a low frequency the auxiliary carrier is obtained by applying the output voltages of two oscillators O_1 and O_2, having frequencies of 9 and 11 Mc/s, to a mixing tube M. The 9 Mc/s oscillator is modulated with the aid of a reactance tube FM. The auxiliary carrier is fed, via filter F and amplifier A_{SP1}, to the part of the transmitter built into the reflector assembly. Here the auxiliary carrier is once more amplified, by amplifier A_{SP2}, and subsequently fed to the oscillating transmitting tube O_3. This is a multi-reflex klystron,

operating at a frequency of about 3500 Mc/s with an output of 10 W. Two stabilizers St_1 and St_2 respectively ensure a sufficient stability of the auxiliary carrier frequency and of the master carrier frequency. The stabilizer is dealt with in the third article of this series [15]).

which is applied via the cable to the other section of the receiver. This contains firstly an I.F. amplifier with a low-noise input stage. The bandwidth of this amplifier (the frequency band in which the amplification does not drop more than 3 dB below that at the central frequency) amounts to 14 Mc/s. In this

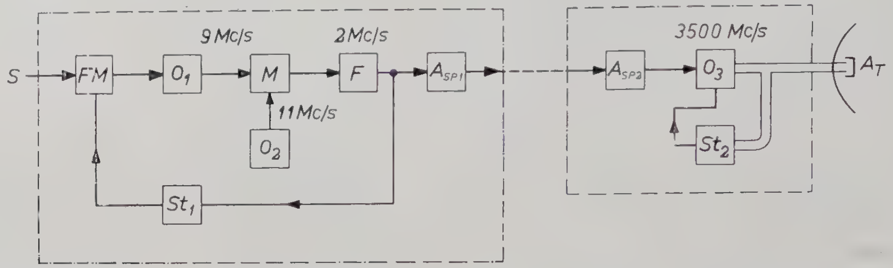

Fig. 8. Block diagram of the transmitter of fig. 7. The left-hand part, surrounded by the dotted line, is built into the rack; the right-hand part, is built into the reflector assembly.

The transmitting antenna A_T consists of a reflector in the shape of a paraboloid of revolution containing, at the focus, a radiator of a type designed by Cutler.

Fig. 9 shows the complete receiver, which is represented by the block diagram of *fig. 10*. The antenna A_R is identical to that of the transmitter. The part of the circuit built into the reflector assembly contains an oscillator O consisting of a triode, the frequency of which is determined by a cavity resonator. In the mixing stage M the antenna signal together with the oscillator signal form an I.F. voltage with a frequency of 58 Mc/s,

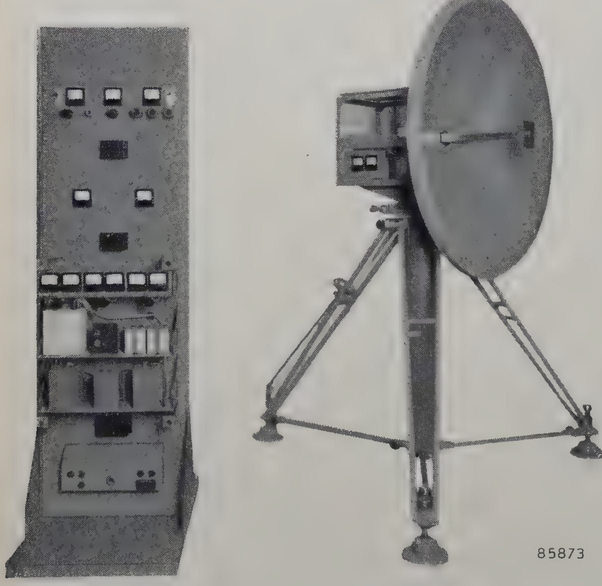

Fig. 9. The experimental receiver for double frequency modulation. To the left is the rack, to the right, mounted on a tripod stand, the reflector assembly containing the first stage of the receiver.

amplifier, staggered circuits are employed. The I.F. signal is applied, via limiter L_M, to the first frequency demodulator D_1 which reproduces the auxiliary carrier with a frequency of 2 Mc/s. This is applied, via amplifier A_{SP}, limiter L_{SP}, and filter F, to the second frequency demodulator D_2, which produces the transmitted signal S.

Measurements on the installation

With the installation in question an experimental telephony link bridging a distance of 12.5 km was established. In view of this fairly short distance, small paraboloid reflectors were used, with a diameter of only 1 m. In order to approximate more closely to the conditions of a link of longer distance, the transmitter output was reduced to 4 W, and an attenuation of 5 dB was incorporated before the I.F. amplifier of the receiver, so that the noise factor of the receiver increased by 4 dB. Because of an obstacle between transmitter and receiver, the signal voltage ultimately received underwent an additional attenuation by 10 dB. The frequency swing with which the carrier was modulated by the auxiliary carrier was 5 Mc/s.

It can be shown that in these experiments, the fluctuation noise was equal to that which would occur with a 50 km link when the received signal is attenuated 24 dB by fading.

If the distance between transmitter and receiver were 50 km instead of 12.5 km, the voltage obtained at the receiver antenna would be four times lower, and therefore reduced 12 dB with respect to the actual conditions used. It should be kept in mind, however, that for such a long-distance link reflectors of larger diameter are used, e.g. 3 m, giving a received signal greater by 18 dB. A further amplification (4 dB) will be obtained from the fact that the transmitter power will be adjusted at 10 W and also from the fact that as a rule there is

[15]) J. Cayzac, Automatic frequency stabilization for a beam transmitter working on centimetric waves, Philips tech. Rev. this issue, p. 334.

no obstacle between transmitter and receiver. Finally, no attenuation will be incorporated before the I.F. amplifier, so that the noise factor of the receiver decreases 4 dB. The ratio of signal to fluctuation noise will thus altogether be $(18 + 4 + 10 + 4) - 12 = 24$ dB higher than the one occurring in the experiment. A fading of 24 dB on the 50 km link therefore makes the fluctuation noise equal to that in the experimental link.

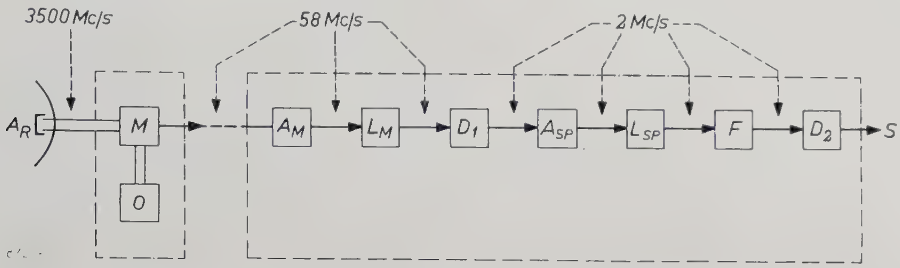

Fig. 10. Block diagram of the receiver of fig. 9. The left-hand section (surrounded by dotted line) is built into the reflector assembly; the right-hand part is incorporated in the rack.

The installation was connected to a carrier-telephony system for 48 channels, of the type described in an earlier issue of this Review [16]). In this system the 48 channels are divided into 4 groups, situated in the frequency bands 12-60, 60-108, 108-156 and 156-204 kc/s. For measuring purposes, incoherent signals supplied by a noise generator were substituted for the speech signal voltages. As a noise generator, use was made of an amplifier of the type applied in the above-mentioned carrier-telephony system for the amplification of a group of 12 channels. The noise voltage supplied by this amplifier was fed, after frequency transformation, to the four groups of channels, 8 channels being kept free. Of each of these "free channels" the noise power W_b was now measured as a function of the power W_s applied to the other channels. In

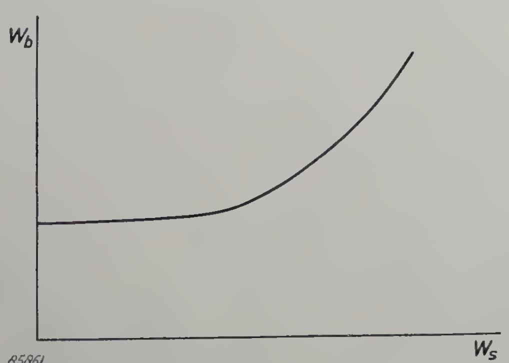

Fig. 11. Noise power W_b in one of the channels, plotted as a function of the power W_s supplied to the other channels. The curve applies to one specific ratio between the level of the signal to be transmitted and the frequency swing.

16) G. H. Bast, D. Goedhart and J. F. Schouten, A 48-channel telephone system, Philips tech. Rev. **9**, 161-170, 1947 and **10**, 353-362, 1948.

this way a curve as plotted in *fig. 11* was obtained. At small values of W_s, W_b consists mainly of the fluctuation noise, which is independent of W_s. At greater values of W_s, however, the intermodulation noise begins to play a part, and W_b increases accordingly.

We have already mentioned in the introduction that, without essentially altering the system, the frequency swing on the auxiliary carrier can be adjusted to various values by incorporating attenuators before the first modulator in the transmitter and after the second demodulator in the receiver, the attenuation of which is varied in the opposite sense. If the modulating voltage is reduced and (consequently also the frequency swing) the fluctuation noise will increase, but the intermodulation noise will become smaller.

When the frequency swing is widened, the reverse will obviously take place. In *fig. 12*, W_b is plotted

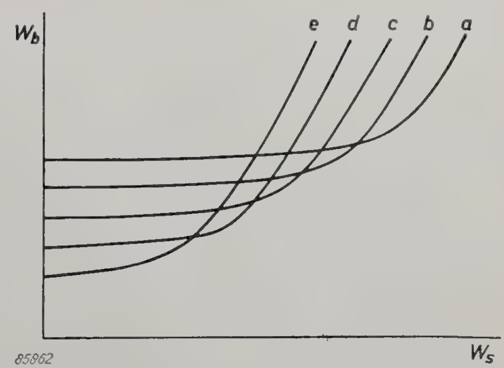

Fig. 12. Noise power W_b in one of the channels, plotted as a function of the power W_s supplied to the other channels, for various values of the frequency swing. For a given signal level, the frequency swing increases from *a* to *e*.

as a function of W_s, for various values of the frequency swing, increasing from *a* to *e*. From this we see that each value of W_s has an appropriate frequency swing producing a minimum value of W_b. In order to adjust the frequency swing to its optimum value, we must know what value of W_s corresponds to the normal power in a 48-channel system. This

power is of course subject to considerable fluctuations. Statistical studies have revealed that with a 48-channel system, the power at points at relative zero level exceeds 1.2 dBm_0 for 1% of the time at the most.

This figure has been derived from an article by Holbrook and Dixon [17]). More recent research, however, has shown that the power values found by them as the normal load of a carrier-wave system can be reduced by 5 dB. This latter consideration has already been allowed for in the above.

We must therefore adjust the frequency swing, by means of the 2 attenuators, so that the minimum value of W_b occurs at this value of W_s. This position was determined experimentally and retained during the subsequent measurements. Curve *a* in *fig. 13*

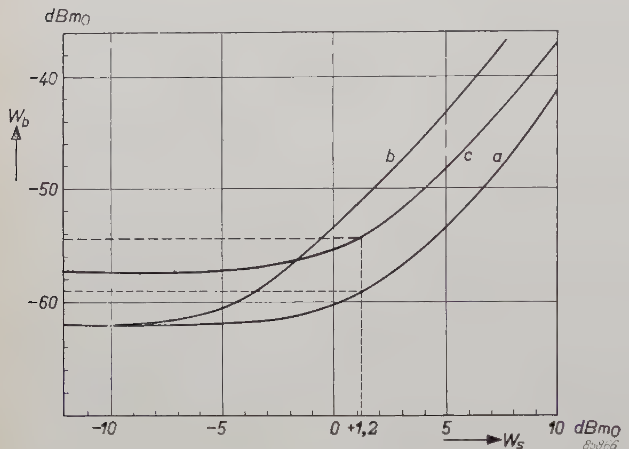

Fig. 13. *a*) Experimental curve, showing the noise power W_b in one of the 8 free channels as a function of the power W_s supplied to the remaining 40 channels.
b) Curve computed from *a* which would apply if, using the same type of transmitters and receivers as in *a*, transmission were over a standard circuit of length 2500 km, the frequency swing being adjusted to the optimum value for each individual link. *c*) Curve which applies if the frequency swing in each of the links were adjusted to the value most favourable to the whole. The power values are plotted in dBm_0, i.e. in dB relative to a power of 1 mW at points at relative zero level.

shows W_b plotted as a function of W_s for one of the channels, viz. that of the highest frequency band (200—204 kc/s). The scale is in dB relative to 1 mW at points at relative zero level, i.e. in dBm_0. The particular frequency swing, earlier designated by *d*, which is representative of a power of 1 mW at points at relative zero level, will consequently occur at $W_s = 0$ dBm_0. From measurements in various channels an average value of 40 kc/s was found for *d*.

Curve *a* in fig. 13 shows that $W_b = -59$ dBm_0 when $W_s = 1.2$ dBm_0 in the channel in question. When W_s is very small, so that we are concerned only with the fluctuation noise, $W_b = -62$ dBm_0. Both as regards fluctuation noise and intermodulation noise, this channel, having the highest frequency band, shows the least favourable properties. (The curves measured in the other channels, however, do not differ a great deal from fig. 13.) Our further discussion can, therefore, be safely restricted to this channel.

Conclusions from the measurements

From the results of the measurements described above, certain conclusions may be drawn concerning the practical value of transmitters and receivers of the type examined in the specified standard circuit, in particular as regards the question whether the C.C.I.F. noise specifications can be complied with. The length of 2500 km of the standard circuit will obviously have to be bridged by means of a number of relay stations. Let us assume that fifty links, each of 50 km length, will be used.

A reduction of the distance between the relay stations and a correspondingly increased number of links, has a favourable effect on the fluctuation noise, since in each link the level of this noise is proportional to the square of the distance, whereas the number of links is inversely proportional to this distance. An increase in the number of relay stations, on the other hand, incurs great expense. For this reason a length of 50 km per link was considered to be a good compromise.

Let us further assume that in two of these links fading will reduce the signal strength to a hundredth (— 20 dB) of its normal value [18]). This means that the fluctuation noise for the total circuit will be $2 \times 100 + 48 \times 1 = 248$ times (24 dB) greater than that of a single link in which no fading occurs. Earlier in this article it was calculated that, in the experiments, the fluctuation noise relative to the signal is 24 dB greater than would occur in a normal 50 km link, so that, as regards fluctuation noise, the experiments are representative of a standard circuit under the conditions mentioned above.

In the standard circuit as laid down by the C.C.I.F., the auxiliary carrier is to be demodulated and then generated and modulated again at nine places. If we presume that the intermodulation noise is formed exclusively in the first modulators of the transmitters and in the second demodulators of the receivers (disregarding the phase distortion in networks and transmission lines), this means that the intermodulation noise will be nine times (9.5 dB) greater than in the experimental link.

[17]) B. D. Holbrook and J. T. Dixon, Load rating theory for multichannel amplifiers, Bell System tech. J. **18**, 624-644, 1939.

[18]) Such a strong fading occurring on two links simultaneously is most unlikely. Measurements on an American-built 107-link beam transmitter chain have shown that a fading of 20 dB occurred on the least favourable of the links during 1% of the time only.

With these data and with the aid of curve *a* of fig. 13, a curve can be plotted, which applies to the whole circuit. This curve, indicated by *b*, would apply if the measurements in question were carried out on the whole standard circuit, in which the frequency swing *d* for each link is adjusted to the optimum value found for a single link. Compared with the experimental link, however, the fluctuation noise and the intermodulation noise have not increased in the same proportion, so that in fact another value of *d* will give better results. In equation (1) A will have retained the same value, but B will be nine times greater. The optimum frequency swing d_{opt} will consequently, according to (2), have to be decreased by a factor of

$$\sqrt[4]{\frac{9}{1}} = 1.73 \ \ (4.8 \text{ dB}).$$

The result is that curve *b* of fig. 13 is shifted upwards and to the right over distances of 4.8 dB, thus forming curve *c*. It can be seen that at $W_s = 1.2$ dBm$_0$, the value of W_b becomes -54.5 dBm$_0$. When the measurement is carried out psophometrically, this value can again be reduced 3 dB, so that we arrive at -57.5 dBm$_0$. The permissible value, as mentioned in the introduction, is 7500 pW, i.e. -51.2 dBm$_0$. It may, therefore, be concluded that even under conditions of severe fading, such as have been presumed for this case, the noise will remain a good 6 dB below the permissible level.

From the measurements of the experimental link it may also be deduced that this level is not exceeded even if the number of channels transmitted is raised to 60; the total noise then increases only by 3.7 dB.

Experiences with the experimental link have furthermore shown that various quantities could be allowed to deviate considerably from their optimum values before the signal-to-noise ratio deteriorated appreciably [19]). The local oscillator of the receiver, for example, can be detuned by as much as 5 Mc/s before the noise rises by 3 dB. (The frequency stability of the oscillator is such that deviations greater then 1 Mc/s are most unlikely.) The central frequency of the auxiliary carrier can be altered by 50 kc/s without more than a 3 dB rise in noise (the maximum deviation from this frequency is restricted to 10 kc/s by the stabilizing circuit). As a general conclusion it may be claimed that the system of double modulation, in spite of its apparent complexity, is characterized by a very high degree of stability.

[19]) These experiments were carried out with the installation connected to a 24-channel carrier telephony system.

In conclusion, the author wishes to convey his thanks to Prof. G. A. Boutry, director of the Laboratoires d'Electronique et de Physique appliquées, for permission to publish this article, and to acknowledge the valuable contribution to this work of G. Andrieux, J. Cayzac, R. Roy and their co-workers, who built the equipment for this experimental link.

Summary. This article describes an experimental beam transmitter and receiver operating on the principle of double modulation. Some of the general criteria for a beam-transmission system for multi-channel carrier telephony are first discussed. In the transmitter, either an amplifying or an oscillating transmitting tube can be employed. Some of the drawbacks of each system are mentioned. When an oscillating transmitting tube is used, the signal to be transmitted has to be directly modulated on this tube. It is very difficult to attain a sufficiently linear frequency modulation at the very high oscillation frequency of this tube. Insufficiently linear modulation causes an excessive intermodulation noise. With *double* frequency modulation, the signal to be transmitted is modulated upon an auxiliary carrier of moderate frequency, so that the requirement of linearity can be easily satisfied. The carrier to be transmitted is then modulated with the auxiliary carrier, a process for which the linearity requirement can be considerably relaxed. Demodulation of the auxiliary carrier need take place only at relay stations where this is required by the telephone service. Many of the relay stations, therefore, contribute hardly at all to the intermodulation noise. An experimental transmitter and receiver was built with the aid of which a telephone circuit with 48 channels was established. Measurements were made of the signal-to-noise ratio. From the results it can be shown that with this equipment, a standard circuit with a length of 2500 km can be built up which satisfies the signal-to-noise requirements laid down by the C.C.I.F.

THE MULTI-REFLEX KLYSTRON AS A TRANSMITTING VALVE IN BEAM TRANSMITTERS

by F. COETERIER.

621.373.423

In recent years there has been a marked increase in the number of types of microwave valves whose operation is based on the transit time effects of electrons. Klystrons, reflex klystrons, travelling-wave tubes and magnetrons are already widely used, while other types are still in a more experimental stage. Each valve has its own particular field of application. The multi-reflex klystron, which is the simplest in design of all the members of the family of velocity-modulated valves, has fully proved its value as a transmitting valve in beam transmission links.

The preceding article [1]) discusses a beam transmitter in which a multi-reflex klystron is used as a transmitting valve. The following characteristics of the valve are important in this application:

a) Relatively high power output at short wavelengths — 10 W or more at 8.5 cm.

b) Possibility of linear frequency modulation over wide bandwidths, without appreciably influencing the power output. A favourable characteristic is that the modulation may be effected by varying the voltage of one or more electrodes which consume hardly any power.

A paper on the multi-reflex klystron has already appeared in this Review [2]); it will be referred to in the following as I. The present paper will report on certain new aspects and deal with the electronic tuning of the valve, a subject which was not considered in I. In conclusion, a description will be given of the construction of the valve.

Principle of the multi-reflex klystron

We shall first recapitulate the operation of the multi-reflex klystron, which was dealt with in detail in I.

The construction is represented diagrammatically in *fig. 1a*. K is the cathode. A and A' are grids forming the anode system. M and M' are grids which are connected to a resonant system and between which a high-frequency alternating voltage is set up.

We shall call the space between M and M' the resonator gap. R and R' are repeller electrodes, which will be dealt with in more detail later. The potential distribution in the absence of a high frequency alternating voltage, is shown in fig. 1*b*; to a first approximation it can be represented by a parabola.

The electrons emitted by K are accelerated and modulated in velocity between M and M' by the high-frequency voltage. Since R is at cathode potential, the retarded electrons turn back before they reach R, while the accelerated electrons penetrate beyond it. The parabolic field between K and R has the property of making the period of oscillation of the electrons constant and hence independent of their velocity. If the field were to remain parabolic between R and R', all electrons would return after the same time interval; in other words, there would be no density modulation of the beam. By arranging for the field between R and R' to deviate from the parabolic form, the transit time of the accelerated electrons is made to differ from that of the retarded electrons, so that, in the returning beam, density modulation is obtained which, by induction between M and M', maintains the high frequency oscillation. The beam then proceeds to the cathode into which all those electrons vanish which have drawn energy

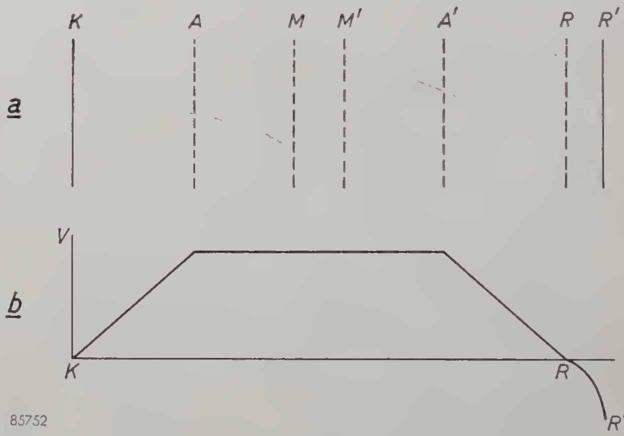

Fig. 1. *a*) Diagram showing the principle of the multi-reflex klystron. K cathode; A and A' grids forming the anode; M and M' grids of the modulator system; R and R' repeller electrodes.
b) Static potential variation between the electrodes.

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[1]) C. Ducot, Beam transmitters with double frequency modulation, Philips tech. Rev., this issue, p. 317.

[2]) F. Coeterier, The multi-reflection tube, a new oscillator for very short waves, Philips tech. Rev. 8, 257-266, 1946, further referred to here as I.

from the alternating field during their outward and return excursion, while the remainder continue to swing to and fro for a certain time between K and R. As the periodic time of the electrons is independent of their velocity, it is possible to adjust the voltage such that the electrons always arrive at the resonator gap (MM') at the right moment, where they repeatedly give up energy. As a result of this periodic surrender of energy, the multi-reflex klystron has a high efficiency as compared with an ordinary reflex klystron (see I). In the latter tube the modulated electron beam returns only once to the resonator gap.

After this introduction we shall now proceed to consider in more detail the operation of the repeller electrodes.

The field between the repeller electrodes

The manner in which the most efficient field might be chosen between the repeller electrodes R and R' was discussed in I. In the meantime, new theoretical considerations have led to a modification which has an important bearing on the construction of the tubes. We shall discuss this theory with reference to *fig. 2*.

It should be noted in the first place that a high-frequency field is present between the grids A and A'. The transit time of an electron between A and A' is not small with respect to one cycle of this field, so that it is difficult to describe the behaviour of the electrons. It appears, however, that the operation of this system is analogous to a similar system in which a high-frequency field is present only in the resonator gap, the remaining space being field-free. We therefore assume that a high-frequency field is present only between M and M', and that this space is so narrow that we can regard the moment of an electron's arrival and departure as a single instant of time t_0. The field outside the resonator gap is independent of the time. The transit time in this field can therefore depend only upon the velocity of the electron.

In fig. 2*a* the high-frequency alternating voltage $v = V_0 \cos \omega t_0$ is shown as a function of ωt_0. According to our assumptions, the change in the energy of an electron which passes the gap at the time t_0 is equal to ev, where e is the electronic charge[3]). It must be remembered, however, that the electrons come from both sides; we assume that an electron coming from the cathode undergoes a change in energy according to fig. 2*a*, while the change in energy of an electron coming simultaneously from the other side will be the inverse of this. If the electrons come from the cathode, we see that

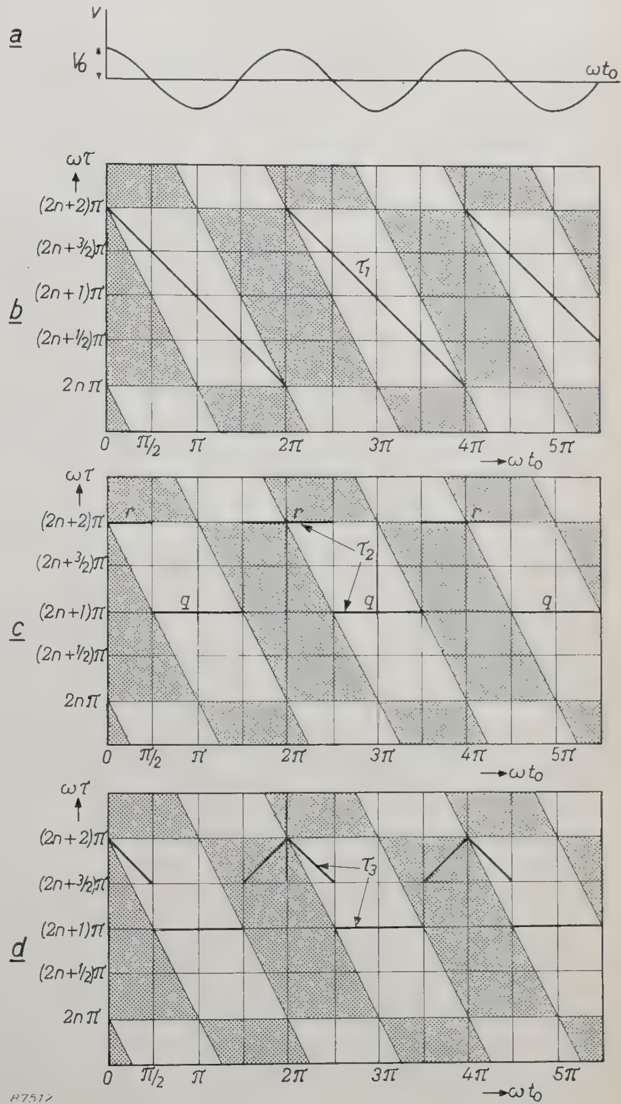

Fig. 2. *a*) The high-frequency voltage $v = V_0 \cos \omega t_0$. *b*), *c*) and *d*) Electron transit-time diagrams; t_0 is the moment at which the electrons first pass through the resonator gap; τ is the transit time, i.e. the time between two outward excursions by the electrons. The various transit-time functions τ_1 (*b*), τ_2 (*c*) and τ_3 (*d*) are described in the text.

they are retarded if $\pi/2 < \omega t_0 < 3\pi/2$, $5\pi/2 < \omega t_0 < 7\pi/2$, and so on. During the intervals these electrons are accelerated. Now we wish to know how great the change is in the energy of an electron after it has travelled once to and fro. If τ is the transit time, i.e. the time between the two successive flights through the resonator gap, the change in energy is evidently

$$\Delta E = eV_0 \cos \omega t_0 - eV_0 \cos \omega(t_0 + \tau)$$

[3]) As a consequence of certain effects such as transit time, space-charge and finite distance of the electron to the boundaries of the grid structure, the change in energy is actually somewhat less than ev, viz. βev, where $\beta<1$. For the sake of simplicity this β-factor is omitted from the argument.

(the second term has a minus sign, because the electron returns in the same field). This expression may be re-written:

$$\Delta E = 2eV_0 \sin \frac{\omega\tau}{2} \sin \omega \left(t_0 + \frac{\tau}{2}\right).$$

If $\Delta E > 0$, the electron has drawn energy from the system. With the aid of the above expression, the areas where $\Delta E > 0$, i.e. the "unfavourable" areas, are shown shaded in the (t_0, τ) plane in fig. 2*b*, 2*c* and 2*d*. The question now arises, how must we choose the transit time τ as a function of t_0 in order to cause the electrons to surrender as much of their energy as possible? For this optimum transit time τ_1 the following expression is valid:

$$\omega\tau_1 + \omega t_0 = 2n\pi \quad (n \text{ an integer})$$

since all these electrons return at the instant $2n\pi$, and in fig. 2*a* we see that the change in energy for returning electrons is then at a negative maximum. The function τ_1 (t_0) is represented in fig. 2*b* by a sawtooth line, and the whole of this line passes through the unshaded region: hence all electrons give up the maximum amount of energy.

The static field needed to produce a transit time function of this nature cannot, however, be realized in practice, since all electrons with the same velocity must have the same transit time and, as fig. 2*a* shows, therefore for all integral values of n we must have for any arbitrary quantity a:

$$\tau\,(n\pi + a) = \tau\,(n\pi - a).$$

The transit-time curve must therefore in any case be symmetrical about the points $t_0 = n\pi$, which is not the case with the sawtooth line. However, this is not the only condition which the transit-time function should satisfy. The transit time of the electrons which are delayed during the first outward movement and which we wish to cause to travel to and fro as frequently as possible, must, moreover, be independent of the velocity, as otherwise their phase would be shifted during repeated reflection. The favourable value for this transit time is thus an uneven number of times π, since every time the electrons return to the resonator they are subjected to an opposing field. Hence.

$$\omega\tau = (2k+1)\pi \text{ for } \frac{(4n-3)\pi}{2} < \omega t_0 < \frac{(4n-1)\pi}{2}.$$

In fig. 2*c* this is represented by the lines q. We still have a certain freedom in our choice of τ in the interlying t_0-regions, (i.e. the times during which the electrons are accelerated when first passing through the resonator gap). The electrons may, owing to their surplus of energy, pass beyond the repeller R (fig. 1), which is at cathode potential, and thus arrive in the space between R and R'. The transit time can be adjusted by the choice of the field in this space. This has no influence on the retarded electrons, because, of course, these turn back before they reach R.

In article I, an optimum transit time was given for the accelerated electrons, which was a constant and was exactly half a period longer than for the retarded electrons (lines r in fig. 2*c*), so that the total transit-time curve τ_2 was a square-wave function. As the figure shows, these electrons after travelling once to and fro, have an energy surplus of exactly zero and were therefore retarded during their return. In so far as they do not arrive at the cathode, they thereafter remain in the same field as the retarded electrons, where their transit time is such that each time they pass through they surrender energy to the resonator circuit.

Efforts have been made in practice to approximate to the transit-time curve τ_2 as closely as possible by means of the static field. The calculation shows that in view of the discontinuities of slope in this curve, there must be a sudden irregularity in the field strength, and this, of course, can be approached only very roughly. It was found, however, that endeavours to improve the slope of this field had the effect of slightly lowering the efficiency of the tubes.

This may be explained as follows. Consider a transit-time curve τ_3 which differs from τ_2 only in the region of the accelerated electrons and also has the necessary symmetry. This curve is drawn in fig. 2*d*. In this case the accelerated electrons fall into 2 groups, depending on t_0. For $\omega t_0 < 2n\pi$ they have a surplus of energy after travelling once to and fro, so that they return to the cathode with finite velocity and thus no longer return to the resonator. For $\omega t_0 > 2n\pi$ they have surrendered energy to the system so that they reverse their direction before reaching the cathode. Owing to the fact that, in this region, curve τ_3 coincides with τ_1, these electrons henceforth always arrive at the favourable moment, that is to say at the moment of maximum inverse voltage in the resonator gap. Upon repeated reflection, a transit time according to curve τ_3 will cause the electrons to give up a greater amount of energy and thus increase the efficiency over that of curve τ_2.

The potential slope corresponding to a given transit-time curve can be found by graphical integration. If we approximate to τ_3 by a somewhat

smoothed curve (which will occur in practice), the potential curve will be as shown in *fig. 3* (curve *c*). This field can be realized fairly well by giving the electrode *R* a specific thickness, thereby realizing the discontinuity of slope in the curve. On the other

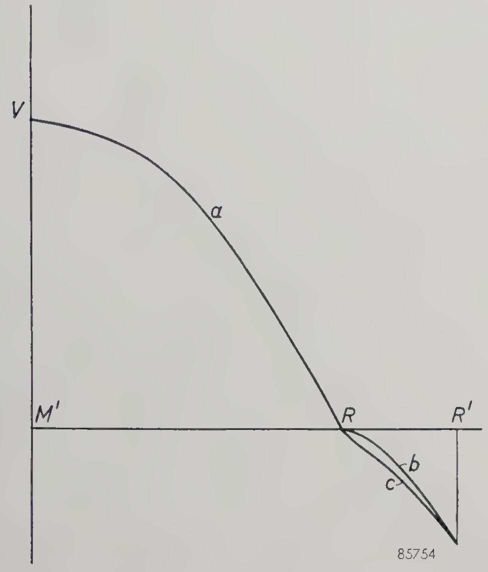

Fig. 3. The static potential between the planes M' and R' of fig. 1. *a*) the ideal potential between M' and R: a parabolic curve. *b*) and *c*) The potential slopes between R and R' corresponding to the transit time curves τ_2 and τ_3 respectively.

hand, the field corresponding to τ_2 (curve *b*) is much more difficult to realize. Thus, a simpler construction results here in better efficiency.

The electronic bandwidth

As stated in the introduction, the frequency can be varied by varying the voltage on one of the electrodes. As the transit time plays an essential part, the valve reaches optimum oscillation and hence delivers maximum power at a specific voltage. The

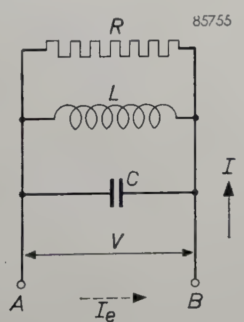

Fig. 4. Equivalent circuit of the high frequency section. *I* high-frequency current in the resonator. I_e high-frequency component of the total electron stream.

difference between the frequency settings at which the output power falls to half its maximum value is called the electronic bandwidth. The equivalent circuit in *fig. 4* may be used for determining this width. Points *A* and *B* represent the openings in the grids *M* and M' of fig. 1. Between these points the following elements are shunted:

1) The resonator, represented by an *LC* circuit. The resonant frequency ω_0 is such that

$$\omega^2{}_0 LC = 1 \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

2) The load, formed by the transformed impedance of the aerial and by the losses in the system. Together, they can be represented by a damping resistor *R* between *A* and *B*. If *V* represents the peak value of the alternating voltage between *A* and *B*, the total high-frequency output *W* is

$$W = \tfrac{1}{2}\frac{V^2}{R} \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (2)$$

The useful output is a certain constant fraction of this.

3) The electron beam, which itself also represents a certain impedance R_e between *A* and *B*.

If, in the circuit formed by these elements, an oscillation is present with a frequency ω and a constant amplitude, the total admittance between *A* and *B* must be zero. We can therefore write:

$$\frac{1}{R}+\frac{1}{R_e}+j\left(\omega C-\frac{1}{\omega L}\right)=0.$$

Assuming that $\omega = \omega_0 + \Delta\omega$, then for $\Delta\omega/\omega \ll 1$ we may write:

$$\omega C - \frac{1}{\omega L} = 2C\Delta\omega$$

so that:

$$\frac{1}{R_e} = -\frac{1}{R} - j\,2C\,\Delta\omega \quad . \quad . \quad . \quad . \quad (3)$$

If I_e represents the beam current component of angular frequency ω, then $1/R_e = I_e/V$. This quantity is called the electronic admittance. We see that the angular frequency ω_0 is generated ($\Delta\omega = 0$) when I_e and V are in phase, and that the frequency is changed when there is a phase shift between them. This phase shift may be brought about by varying the transit time of the electrons in the static field, which is done in practice by changing the potential on electrode *R* (fig. 1). The relation between this potential and the frequency is called the modulation characteristic.

In principle, it would be more correct to modulate the electrodes *A*, A', *M* and M' (fig. 1) together, since by modulating *R*, the transit time is changed on only one side of the resonator. In practice, however, it is important for the modul-

ation to consume as little power as possible, so that it is preferably carried out on those electrodes over which hardly any current flows. Owing to the asymmetrical transit time, the bandwidth is slightly reduced, an effect which is tolerated for the reasons given.

The question is now how large $\Delta\omega$ can be without causing a severe depreciation of the high-frequency output. To determine this exactly, it would be necessary to know the relation between I_e and V explicitly at a given transit time. This rather elaborate calculation exceeds the scope of this paper, but we can say something in general terms about $\Delta\omega$. It is evident that when I_e is shifted 90° in phase with respect to V, there can be no more energy supplied. To a rough approximation, therefore, a 45° phase shift means that there will still be half the output left, and this in fact roughly agrees with the theoretical results [4]).

Taking $\Delta\omega_r$ as the frequency variation occurring at a 45° phase shift, then according to equation (3):

$$\frac{1}{R} \approx 2C\,\Delta\omega_r$$

and the total electronic bandwidth $= 2\Delta\omega_r \approx 1/RC$. The value of C is exclusively determined by the construction of the resonator. As regards $1/R$, we find from (3) that

$$\frac{1}{R} = \left(\left|\frac{I_e}{V}\right|\right)_{\omega=\omega_0}.$$

This shows the advantage of the multi-reflex klystron over an ordinary klystron; owing to the repeated reflections, the alternating current component I_e can take on much higher values, so that the bandwidth is many times larger.

It is true that bandwidths of the same order of magnitude can be realized with ordinary reflex klystrons, but this is done by keeping the direct voltage very low, as a result of which V remains much smaller. This in turn greatly reduces both the high-frequency output and the efficiency (to no more than a few watts and a few % respectively). Reflex klystrons with this power output can indeed be used as transmitting valves in a link transmitter, but experience shows that the greater power of the multi-reflex klystron is a considerable advantage.

Certain types of high-frequency valves exist, viz. travelling-wave tubes [5]), with which an even greater electronic bandwidth is possible. However, their efficiency is lower than that of the multi-reflex klystron, and their construction presents greater technological difficulties.

We shall now briefly review the average characteristics of the series-manufactured valve used in the link transmitter discussed in the preceding article [6]).

Wavelength	8.5 cm
Anode direct voltage	3000 V
Anode direct current	20 mA
Useful power output	12 W
Efficiency	20 %
Electronic bandwidth	20 Mc/s.

The high frequency voltage V' is measured as follows. The extra energy of the maximally accelerated electrons amounts to eV'; thus V' is the exact potential difference that must exist between R' and K (fig. 1) in order to suppress all current flowing to R'. The potential difference in this case is:

$$V' = 600 \text{ V}.$$

At $\Delta\omega = 0$ the resultant high-frequency output is equal to $\frac{1}{2} I_e V'$. This value, owing to the losses, is greater than the useful output, and in the present example can be taken at 24 W. It therefore follows that:

$$I_e = 80 \text{ mA}.$$

Since the theory shows that I_e would be approximately equal to the anode direct current I_0 if the electrons were to pass through only once, we must have a fourfold reflection. The maximally retarded electrons lose $4 \times 600 = 2400$ eV, that is to say most of their energy. This fourfold reflection is in good agreement with conclusions derived from the modulation characteristic (see [4])).

Construction of the tube

The construction of the multi-reflex klystron, a photograph of which is shown in *fig. 5*, includes elements of conventional transmitting-valve design and is very simple for a high-frequency valve.

The oscillating system M and the anode system A, with which the desired potential slope is adjusted, are connected. (M is not visible in fig. 5). Both are made of molybdenum sheet and the seams are brazed together with copper or nickel. The electron beam passes simply through the holes in the electrodes. By using grids in these holes it would be

[4]) Since I_e is the result of a number of reflections, the change of the transit angle $\omega\tau$ for one movement to and fro is much less than 45°. The amount by which the voltage must be changed in order to bring the valve back to optimum generation accurately indicates the number of reflections occurring.

[5]) J. R. Pierce, Traveling-wave tubes, D. van Nostrand Co. (1950).

[6]) The same valve is also being used in the channel link transmitter for relaying Eurovision television programmes. See: A. Laurens, Onde él. **34**, 999-1005, Dec. 1954.

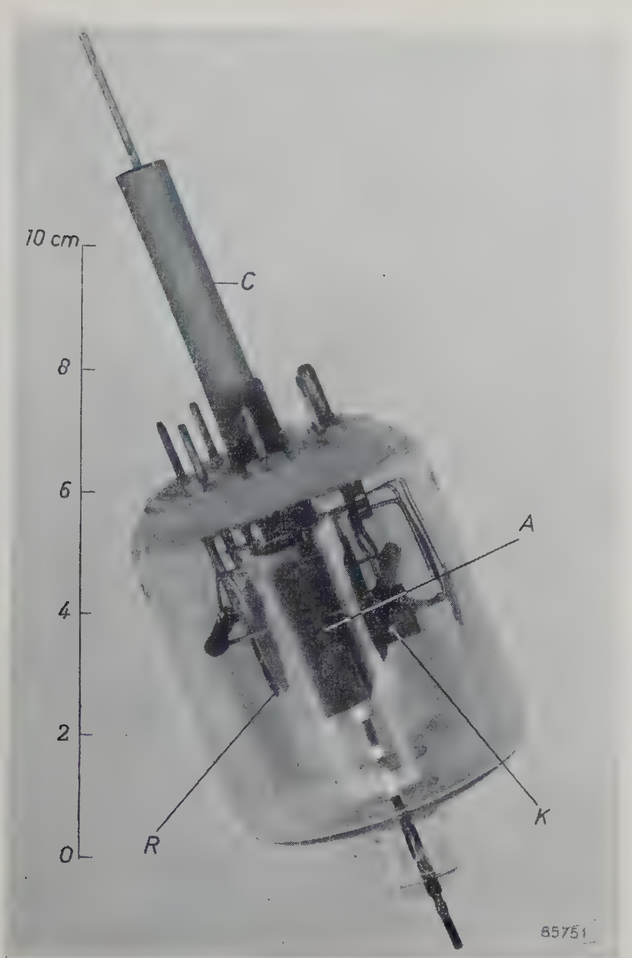

Fig. 5. The multi-reflex klystron used in the beam transmitter. *A* anode system, *C* coaxial output system. *K* cathode. *R* repeller electrodes.

possible to obtain a β factor closer to 1 (see [3]) but the grids would get too hot owing to bombardment by electrons. The correct frequency is pre-set by means of a correction screw in the side-wall of *M*. The connection for the H.T. electrode is arranged at the top of the valve to avoid flash-over and electrolysis of the glass [7]). The glass envelope is made fairly wide to minimize the influence of ambient temperature.

[7]) These and other improvements in design were introduced by E. G. Dorgelo & H. G. Gerlach of the Electronic Tubes Development Laboratory, Eindhoven.

The other components are mounted on a sintered glass base [8]). The external lead of the coaxial output system is formed of molybdenum sheet rounded into a cylinder; a small gap is left between the edges which facilitates the sealing-in. Furthermore, the structure is now more resilient: the valve can now readily be inserted in a coaxial plug-socket. The output system is not connected with the anode, so that it need not be at high tension and the outer sleeve can be earthed.

Since the holes through which the electron beam passes have to be fairly small in order to obtain good interaction of beam and electrodes, the cathode must be able to deliver a current density of approximately 2 A/cm^2 at the given output. The cathode is, moreover, exposed to bombardment by accelerated electrons. No oxide cathode would be able to stand up for long to such conditions. The obvious solution for this valve was therefore to use the L-cathode, already described in this Review [9]), which has a high maximum emission and is proof against electron bombardment.

For concentrating the electrons, an axial magnetic field with an induction of about 700 gauss is needed; this is provided by a "Ticonal" permanent magnet. Valve and magnet together weigh approximately 4 lbs.

[8]) E. G. Dorgelo, Philips tech. Rev. **8**, 2-7, 1946.
[9]) H. J. Lemmens, R. Loosjes and M. J. Jansen, "A new thermionic cathode for heavy loads", Philips tech. Rev., **11**, 341-350, 1949/50.

Summary. The multi-reflex klystron, which was first described in this Review in 1946, is eminently suitable for use as a transmitting valve in beam transmitters, owing to its large electronic bandwidth and its relatively high output power at centimetre wavelengths. The first part of this article develops the theory of the valve. It is shown that the optimum transit time of the electrons as a function of the moment of time at which they pass through the resonator enables the requisite potential curve to be easily established between the repeller electrodes. It is then demonstrated that the large bandwidth of the valve is due to its low electronic admittance, which in its turn is due to the repeated reflection of electrons. From tests made on the valve, it appears that the electrons which surrender energy are reflected four times. As compared with other high-frequency valves, the multi-reflex klystron is very simple in design. Most of its components are made from molybdenum sheet mounted on a sintered glass base.

AUTOMATIC FREQUENCY STABILIZATION FOR A BEAM TRANSMITTER WORKING ON CENTIMETRIC WAVES

by J. CAYZAC *).

621.316.726:621.396.61.029.64

This article, the third relating to a centimetric telecommunication transmitter, describes a device with which the oscillation frequency of the transmitting valve (a multi-reflex klystron) can automatically be kept constant to within 1 in 10 000.

The carrier frequency of the radiation emitted by a beam transmitter working on centimetric waves is required to be extremely stable. To avoid interference between different relay stations, an accuracy of 1 in 10 000 may be necessary at a carrier frequency of about 3400 Mc/s (8.5 cm wavelength). This is especially the case in long-distance communication via a succession of relay stations, the carrier frequencies of which have various values within a certain frequency band.

The preceding article in this Review describes a transmitting valve, the multi-reflex klystron [1]), which is particularly suitable for use in beam transmitters owing to its ability to supply an output of at least 10 W at centimetric wavelengths and because of the fact that it can readily be frequency modulated. The frequency at which this valve oscillates is dependent upon various quantities, especially upon the supply voltages (which, incidentally, is why it can be so easily modulated). At a nominal frequency of 3400 Mc/s, the anode voltage of the valve is 3000 V, and for every volt variation on the anode a frequency shift results of 100 to 200 kc/s. Careful stabilization of the supply voltages would therefore seem to be called for. However, since the frequency may vary for other reasons, because of thermal effects, for instance, it is necessary to find a means of automatically compensating any frequency change that might occur, irrespective of the cause. The power used for this purpose must of course be only a small fraction of the useful power.

The carrier frequency of a transmitter can be kept constant within accurate limits by comparing it with the frequency (many times multiplied) of a crystal oscillator, and by using any difference between them for corrective regulation. This article, however, is concerned with a different procedure which, although less accurate, nevertheless satisfies the requirements and offers, moreover, a number of practical advantages.

The frequency correction device to be described consists of two sections. The H.F. section, which must be introduced between the transmitter valve and the antenna, contains no electronic valves or other elements requiring maintenance. Since relay transmitters are often mounted on high towers, this can be an important consideration. A cable leads to the second (low-frequency) section, in which exclusively conventional receiver valves are used.

Principle of the frequency correction

Consider a symmetrically modulated carrier-wave with a frequency f_c. The spectrum of the radiated power may be represented by a band lying symmetrically about f_c and of width equal to twice the frequency swing, $2f_z$ (*fig. 1a*). A small part of this power is fed to two cavity resonators whose resonant frequencies, f_1 and f_2, lie just outside the radiated spectrum, i.e. the frequency difference $f_1 - f_2$ is taken somewhat larger than $2f_z$. The Q factor of both cavity resonators is so adjusted as to cause their resonance curves to overlap somewhat, as shown in fig. 1*b*. At a given carrier frequency f_c, the average power in each resonator can now be measured with the aid of a crystal detector.

Assuming that the crystals, and their coupling with the cavities, are identical and not selective, the same signal will be obtained from both crystals only when the carrier frequency f_c lies midway between the frequencies f_1 and f_2 (fig. 1*c*) i.e. when it is equal to f_0, or $f_0 = \frac{1}{2}(f_1 + f_2)$. The signals in the two cavity resonators are represented by the hatched areas. If, as in fig. 1*d*, f_c lies closer to f_2 than to f_1, the average signal in cavity *2* will be stronger than in cavity *1*, and vice versa. Alternate measurement of the signal from both cavities results in a current with an A.C. component, the amplitude of which varies according to the difference $|f_c - f_0|$. At $f_c = f_0$, the alternating current is at zero. The sign of $f_c - f_0$ determines the phase. After amplification,

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[1]) F. Coeterier, The multi-reflex klystron as a transmitting valve in beam transmitters, Philips tech. Rev., this issue, p. 328

the alternating current is fed to a phase-sensitive detector circuit to produce a direct voltage, the magnitude and sign of which are proportional to $f_c - f_0$. If this output voltage, after amplification, is applied to an electrode of the valve which controls

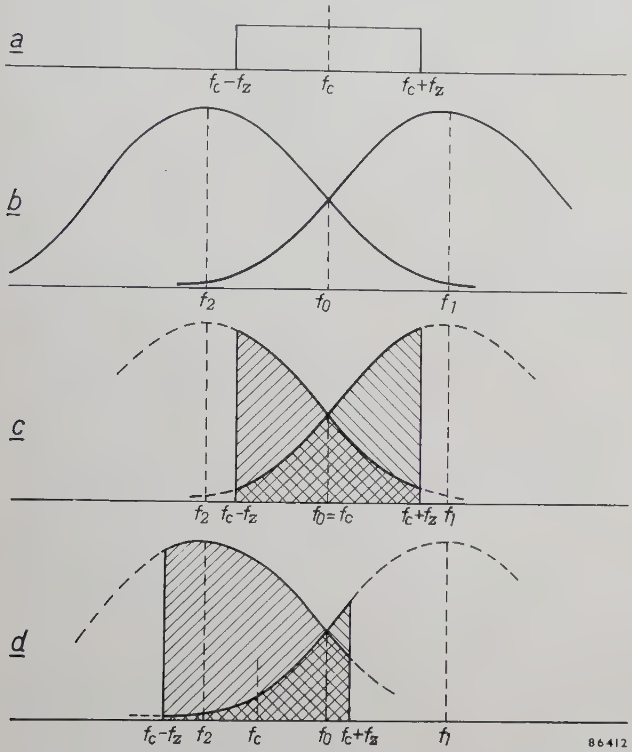

Fig. 1. *a*) The transmitted frequency band, showing the central frequency f_c and frequency swing f_z. The signal strength is plotted vertically.
b) The band-pass curves of cavity resonators *1* and *2*, with resonant frequencies f_1 and f_2.
c) The signals (hatched areas) produced by the frequency band in (*a*) in cavity resonators *1* and *2*, when $f_c = f_0$.
d) The same when $f_c \neq f_0$.

the frequency, it will automatically bring the carrier frequency f_c back to the frequency f_0.

It is essential that the characteristics of both crystals remain constant within very narrow limits over the course of time. Since this condition is difficult to fulfil in practice, we have evolved a method which entirely eliminates differences of this kind. The method involves the use of the same crystal for measuring the fields in the two cavity resonators alternately, in a manner to be described below.

The above arguments are strictly applicable only if the modulation causes the spectral components of the transmitted frequency band to be symmetrically distributed about a fixed frequency. However, the system can still be used if the distribution is asymmetric (as, for instance, in modulation with television signals, when one side band is suppressed). In that case it is possible to design an arrangement with which uses only one signal at a fixed place in the transmitted band. For this purpose the signal can be that which fixes the zero brightness level (black level) of the television picture.

It should be added that the system described is especially suitable for use when the transmitted frequency band is wider than the permissible frequency shift. That is generally the case with telecommunication systems working on decimetric or centimetric waves.

Design of the stabilizer

The system is used on a beam transmitter with a multi-reflex klystron [2]), the output power of which is 10 W at a frequency of 3400 Mc/s. The frequency band produced by modulation amounts to a maximum of 6·5 Mc/s above and below the carrier frequency. By varying the voltage on the repeller electrode, the central frequency can be varied by 50 kc/s per volt in a range from — 100 to + 100 V with respect to the cathode potential.

The high frequency section

The bandwidth required is 13 Mc/s. In our case the resonant frequencies of the two cavity resonators are set at

$$f_1 = f_0 - 7.5 \text{ Mc/s and } f_2 = f_0 + 7.5 \text{ Mc/s}.$$

The whole radiated band thus falls between f_1 and f_2 (as in fig. 1). The relation between the difference of the signals in both resonators and the frequency f_c should be made as linear as possible. This is done by adjusting the Q factor of the cavities to a certain value, which in the present instance was found to be 280.

This value may be derived as follows. As a function of the frequency f, the high-frequency voltage V measured in a cavity resonator is

$$V = \frac{V_0}{1 + 4Q^2\left(\frac{f-f_r}{f_r}\right)^2},$$

in which f_r is the resonant frequency of the cavity, Q the quality factor and V_0 the voltage for $f = f_r$. The characteristic of the discriminator $y(f)$ is given by the difference of the voltages in the two cavities. Assuming that both V_0 and Q are equal in both cavities, we obtain:

$$y = \frac{V_0}{1 + 4Q^2\left(\frac{f-f_1}{f_1}\right)^2} - \frac{V_0}{1 + 4Q^2\left(\frac{f-f_2}{f_2}\right)^2}.$$

If we put

$$2Q\,\frac{f-f_0}{f_0} = x,$$

and

$$2Q\,\frac{f_0-f_1}{f_0} = 2Q\,\frac{f_2-f_0}{f_0} = a,$$

then, since $f_1 \approx f_2 \approx f_0$, it follows that

$$y = V_0\left[\frac{1}{1+(x+a)^2} - \frac{1}{1+(x-a)^2}\right].$$

[2]) C. Ducot, Philips tech. Rev., this issue, p. 317.

At $x = 0$, $d^2y/dx^2 = 0$, so that the characteristic is already fairly linear in the vicinity of $x = 0$. The linearity can be further improved if

$$\left(\frac{d^3y}{dx^3}\right)_{x=0} = 0,$$

which is the case when $a = \sqrt{1.5}$, so that

$$Q = \frac{f_0\sqrt{1.5}}{2(f_0-f_1)}.$$

With the values given for f_0 and f_1, we find $Q = 280$.

At the frequencies concerned, a Q factor of this order can easily be obtained. It is, however, essential that both cavity resonators have the same Q and that both are excited in the same way. For this purpose we have adopted the construction shown in *fig. 2*. The two cavities are milled from a single block of copper and are weakly coupled via identical coaxial conductors to the wave guide through which the high-frequency energy to be radiated passes. Both coupling systems are arranged symmetrically in a plane perpendicular to the direction of propagation, so that the coupling is identical in amplitude and phase. The total power consumed by the systems is about 0.5 mW, which is a mere fraction of the output power of about 10 W.

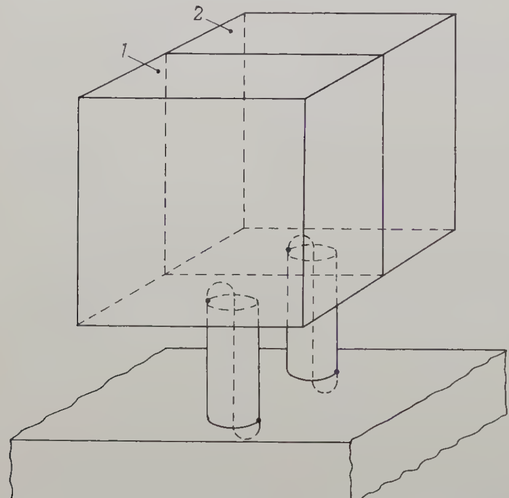

Fig. 2. The two cavity resonators *1* and *2* are symmetrically coupled to the wave guide in a plane perpendicular to the direction of propagation.

Fig 3 illustrates what would happen if each resonator were provided with a crystal detector and if one crystal had a higher sensitivity than the other. The frequency at which the output signals are identical is shifted from f_0 to f_0'. Moreover, even if f_c were now to coincide with f_0', unequal signals would be formed by the modulation in both crystals. As may be seen from fig. 3, the signal in cavity *1* is larger, so that f_c would be further shifted towards the left. The result would therefore be an impermissible frequency shift. For these reasons, the voltage in both cavities is measured alternately with one and the same crystal as stated earlier. This alternation of measurement can be done at a low frequency, because the frequency shifts concerned take place very slowly. The control apparatus can therefore have a time constant of about one second.

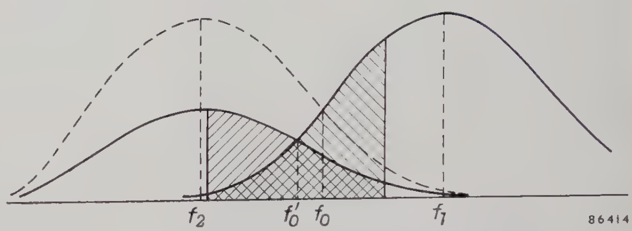

Fig. 3. If the detector in cavity resonator *2* is less sensitive than that in *1* the band-pass characteristic obtained will be that shown by the full-curve instead of the broken curve. Point f_0 is shifted to f_0'.

The coupling of the two resonators with the crystal is shown schematically in *fig. 4*. The crystal *4* is arranged in complete symmetry with respect to the resonators *1* and *2*, and is coupled identically with the resonators through a small hole in the partition wall *3*. The commutation of the crystal is effected by two "absorption modulators" in anti-phase, one in each resonator. In operation, the energy in one resonator is almost completely absorbed, while the other resonator oscillates unattenuated. In the next half cycle, the absorption modulator operates in the other resonator.

These absorption modulators make use of the fact that, at the frequencies concerned, the losses of some ferrites are highly dependent upon a weak, external magnetic field [3]).

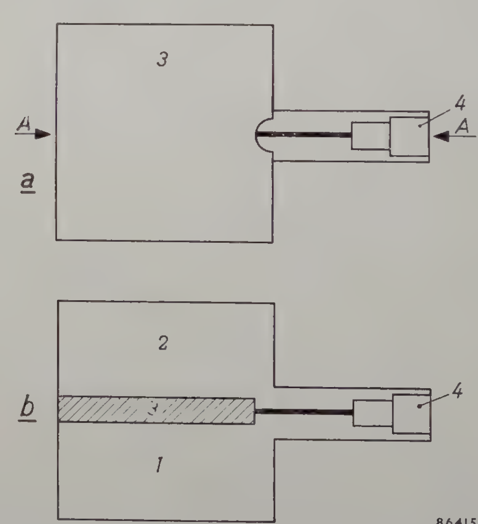

Fig. 4. Positioning of the crystal *4* in the plane of the partition wall *3* between resonators *1* and *2*. *a*) Side view, omitting resonator *1*, *b*) cross-section at *AA*.

[3]) See H. G. Beljers, W. J. van de Lindt and J. J. Went, J. appl. Phys. **22**, 1506, 1951. The phenomenon in question, and its application for modulating the amplitude of microwaves, will shortly be dealt with in this Review.

The system used here is shown in *fig. 5*. Two coaxial lines, K_1 and K_2, which are coupled at one end with the cavity resonators, are terminated at the other end by a loop through which rods of ferroxcube, F_1 and F_2, are passed. Around each rod is a coil, a varying current through which produces a varying magnetic field in the rods. The high-frequency losses in each of the rods vary corresponrods with the aid of permanent magnets (M_1 and M_2 in fig. 5) to a value corresponding to an "average" Q (points A and B in fig. 6). By oppositely premagnetizing the two rods the same current can be made to flow in both coils. As appears from fig. 6 the Q factors of the two resonators are then modulated in opposite senses.

The chief problem here is that the Q factor of

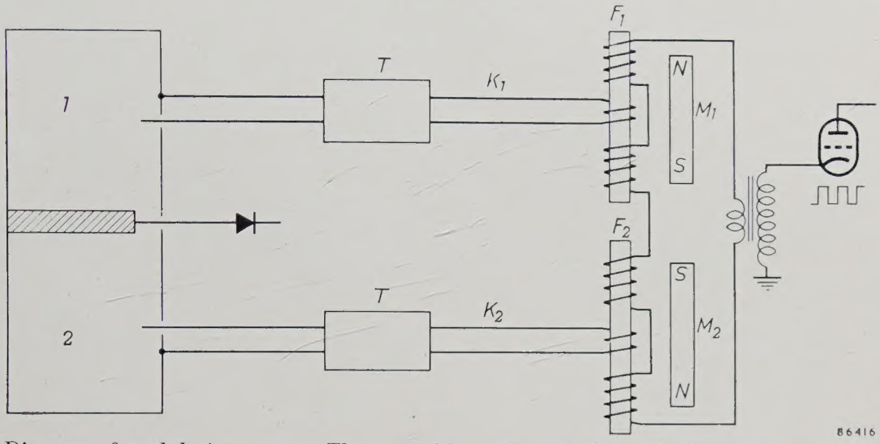

Fig. 5. Diagram of modulation system. The coaxial lines K_1 and K_2 couple the resonators *1* and *2* to the ferroxcube rods F_1 and F_2, which are successively magnetized by an alternating current of rectangular waveform. The permanent magnets M_1 and M_2 pre-magnetize the ferroxcube in opposite directions; T = matching transformers.

dingly and thus vary the Q factor of the cavity resonators via the coaxial lines. The rods are matched to the resonators by impedance transformers T.

Fig. 6 shows how the Q factor of a cavity resonator depends upon the magnetic field in which the ferroxcube is placed. It can be seen that the value of Q is independent of the direction of the field. If, as in our case, we wish to modulate with a rectangular waveform, it will mean that the current fed to the coils will have a DC component. This raises difficulties, since the coils are fed by a matching transformer. It is simpler to premagnetize the ferroxcube

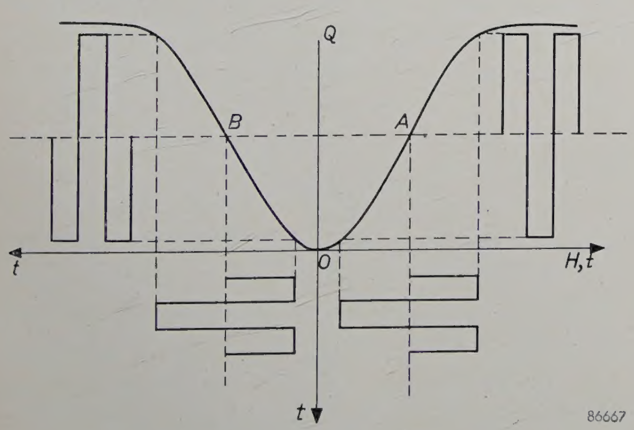

Fig. 6. The quality factor Q of the cavity resonator as a function of the magnetic field H surrounding the ferroxcube. The Q factor of the two cavity resonators is oppositely modulated by oppositely pre-magnetizing the ferroxcube rods, the direction of H being the same.

each resonator in the active phase must always have the same value. The problem is simplified in this case by the fact that, with a magnetic field above a certain strength, the losses in the ferroxcube are fairly constant and very low. Hence by choosing the amplitude of the alternating current on the large side, small changes in amplitude have hardly any effect on the value of Q.

It has been found that variations in the temperature of the different circuit elements coupled to the cavity resonators (fig. 5) give rise to slight variations in the resonant frequencies. For this reason, the whole circuit is placed in a thermostat controlled enclosure.

The low-frequency section

As we have seen, a device is needed that can distinguish from which cavity the stronger signal originates, and which enables the sign of the correction voltage to be determined. This device, a phase discriminator, functions as follows. Two signals, S_1 and S_2, which have the same waveform but are in anti-phase, are applied to the screen grids of two pentodes, connected as shown in *fig. 7*. These signals are derived from the square waveform current S_0, flowing in the coils around the ferroxcube rods. The duration of the (positive) pulse is, however, reduced by one half, as shown in fig. 7. The two control grids are negatively biased such that

both valves are normally cut off, irrespective of the potential on the screen grids. The square-wave control signal from the crystal is applied, after amplification, to both control grids, but only that valve conducts which receives positive pulses simultaneously on its control and screen grid. The amplitude of the anode current is proportional to the amplitude of the alternating voltage on the grids. According to whether resonator *1* or *2* gives the stronger signal, anode current will flow in valve B_1 or B_2.

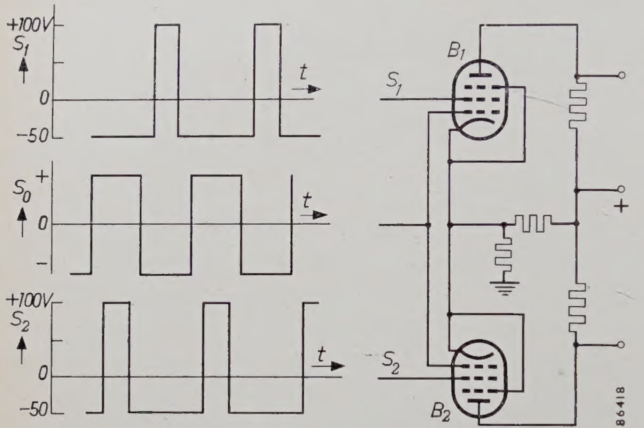

Fig. 7. The phase discriminator. Signals S_1 and S_2, derived from a signal S_0 proportional to the current through the coils, are applied to the screen grids of pentodes B_1 and B_2.

The remaining part of the circuit is shown in *fig. 8*. The output voltages from the pentode are rectified by diodes D_1 and D_2, and the rectified voltages are fed to the grids of two triodes connected as cathode followers. The cathode of one is earthed (C) and that of the other (A) is connected to the repeller electrode of the multi-reflex klystron. The repeller voltage with respect to earth is

$$V_A - V_C = (V_A - V_B) - (V_C - V_B),$$

i.e. equal to the difference of the cathode-grid voltages of both triodes and proportional to the difference between the rectified signals. If the detectors D_1 and D_2 pass no current, this voltage will be zero and the frequency will not be affected, However, as soon as one of the diodes begins to pass current, the repeller bias will change in the direction appropriate to that diode, and the frequency will be automatically corrected until the current returns to zero. The characteristics of the two output valves should of course be fairly similar. It has been found in practice that differences in valve characteristics cause frequency deviations of less than 1 in 10^6 which is considerably less than the deviation of 1 in 10^4 permissible.

The low-frequency section can be designed in another way. The multi-reflex klystron can be tuned not only electrically but also mechanically. By means of a mechanical system connected to the built-in cavity resonator, it is possible to vary the frequency by a total of 300 Mc/s, which is much more than can be achieved by adjusting the repeller potential. For this purpose a servo motor with two windings is used as the phase discriminator. The alternating voltage at the output of the circuit in fig. 7 is fed, after amplification, to one of the two windings. The other winding receives continuously a signal of the same frequency but whose phase differs from that of the first signal by 90°. As we have seen, the phase of the first signal changes by 180° if the sign of the frequency deviation reverses. In this way a rotating field is obtained which causes the rotor of the servo motor to rotate in a direction depending upon the phase of the first signal, and rotation continues until the balance is restored. With this system the frequency can be kept constant with approximately the same accuracy. The great advantage of mechanical compensation is that the repeller bias remains constant. After the latter has been adjusted to obtain the most linear possible modulation of the transmitted carrier wave (see [2]), the linearity is maintained when the compensation system comes into operation.

The stabilization ratio

The control range of the repeller, which is determined by the characteristics if the multi-reflex klystron, is limited to ± 100 V. This includes a large safety margin. The frequency change a per volt change of repeller potential determines the maximum frequency deviations which the system can correct without departing from the permissible voltage zone. In the present equipment, a = 50 kc/s per volt, which means that it is possible to correct frequency deviations up to 5 Mc/s.

Our object was to prevent the occurrence of frequency shifts greater than about 250 kc/s which implies an accuracy somewhat better than 1 in 10^4. The amplifier is so designed as to make full use of the repeller control range at the frequency variations concerned; the maximum shift of 250 kc/s thus gives rise to a correction voltage of 100 V. Without correction, the frequency in this case would have drifted 5 Mc/s. The ratio of the frequency variations with and without correction is called the stabilization ratio: with the present system it amounts to 20.

For a deviation of 250 kc/s, the crystal receives a voltage of about 1 mV. The voltage gain of the phase

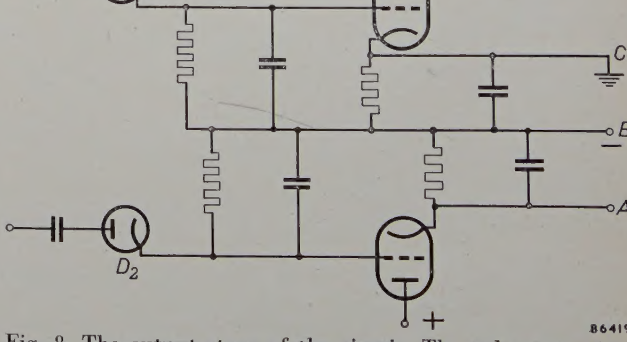

Fig. 8. The output stage of the circuit. The voltage at A is fed to the control electrode of the transmitting valve.

discriminator is 60, so that further amplification by 1700 is needed to obtain the output voltage of 100 V. At the low frequencies in question, however, this presents no difficulties. In practice we usually employ 1.5 times as much amplification in order to allow for any possible decrease in voltage gain.

Adjustment to the desired frequency

In view of the fact that it is sometimes necessary to change the frequency of the transmitter quickly, the adjustment required should be as simple as possible. This means that it must be possible to change the resonant frequencies of both cavity resonators by an equal amount simultaneously without seriously influencing the Q factor or the coupling. The method adopted is to fit each cavity resonator with an adjusting screw, mechanically coupled so that they turn together. With this system a frequency band of 150 Mc/s can be covered, which is the total bandwidth within which the multi-reflex klystron can be mechanically tuned. At the limits of this band the output voltage from the crystal — and hence the stabilization ratio — is 2.5 times lower than in the centre of the band.

The difference is normally unimportant, but should it be disturbing it can be remedied by, for example, coupling the resonators a little more closely to the wave guide. This consideration apart, it is possible with the mechanism employed to change the frequency of the transmitter within a band of 150 Mc/s by a single, brief adjustment.

The H.F. section of the frequency stabilizer with the mechanical tuning system is reproduced in *fig. 9*. The thermostat enclosure is not shown in the photograph. When the apparatus is in operation, the mechanical adjustment can be made externally via a thermally insulating spindle which projects through the enclosure.

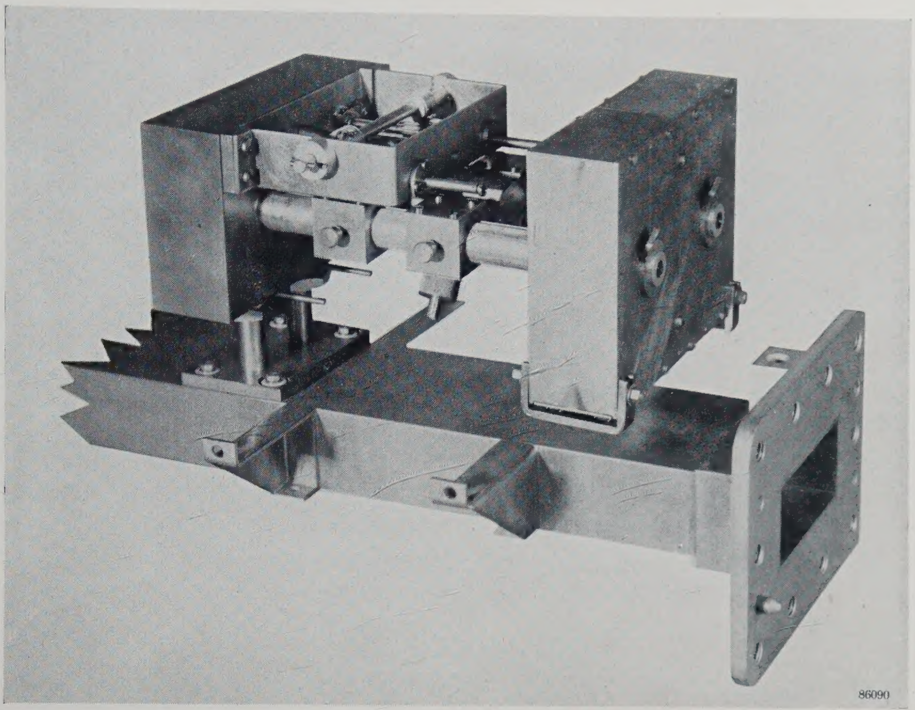

Fig. 9. The H.F. section of the stabilization circuit. The photograph shows the box with the two cavity resonators (left) and the co-axial couplers which lead to the wave guide; two coaxial lines lead to the ferroxcube modulator (right). The mechanical tuning system can partly be seen.

Summary. An automatic stabilizing device is described with which the frequency of a beam transmitter working with a multi-reflex klystron on a wavelength of 8.5 cm, can be kept constant within an accuracy of 1 in 10^4. Two cavity resonators, whose resonant frequencies differ by about 15 Mc/s, but which are otherwise identical, are coupled to the wave guide. If the frequency of the transmitted carrier lies midway between the two resonant frequencies, a slight change in it will produce a stronger signal in one resonator and a weaker signal in the other. The difference signal, whose amplitude and sign are a measure of the frequency shift, is amplified and fed to a control electrode of the transmitting valve, which automaticaly corrects the frequency. The signals in the two cavity resonators are measured alternately by the same crystal detector, so that the measurement is not influenced by small differences (or possible drifting) between the characteristics of two otherwise suitable detectors. The commutation is effected by "absorption modulation" with the aid of small rods of ferroxcube, whose energy absorption depends upon the strength of an alternating magnetic field. The resultant alternating voltage on the crystal is rectified and fed to the control electrode via a phase discriminator. The stabilization ratio (the ratio of the frequency shifts with and without stabilization) is 20. With the multi-reflex klystron, the transmitter frequency can also be corrected by mechanical means; the accuracy within which the frequency can thus be kept constant is comparable to that achieved by purely electronic means. The entire system can be mechanically adjusted over a band of 150 Mc/s.

ABSTRACTS OF RECENT SCIENTIFIC PUBLICATIONS BY THE STAFF OF N.V. PHILIPS' GLOEILAMPENFABRIEKEN

Reprint of these papers not marked with an asterisk * can be obtained free of charge upon application to Philips Electrical Ltd., Century House, Shaftesbury Avenue, London W.C. 2.

2241: L. Heyne: Operation and characteristics of the "Vidicon" (T. Ned. Radiogenootsch. **20**, 1-10, 1955, No. 1).

The operation of "Vidicon" is explained and compared with that of the "C.P.S. Emitron". In both tubes a charge image is generated on the light-sensitive surface by the optical image focussed on it. However the "C.P.S. Emitron" uses photo-emission for this conversion whereas in the "Vidicon", photoconductivity is used. The video signal is obtained in both tubes by scanning the charge image with a beam of low velocity electrons, which stabilize the surface of the target at the potential of the cathode of the electron gun. The photoconductor used in the "Vidicon" must have a very high specific dark resistance, good sensitivity for visible light and no retardation effects.

With lead oxide encouraging results have been obtained. Lead oxide vidicons show a very low dark current, a good sensitivity and can be made with a short decay time. The signal output is proportional to the illumination. The lead oxide layer is also sensitive to X-rays. In X-ray pictures the statistical fluctuation of the number of X-ray quanta is easily noticeable. This is the result of the fact that one X-ray quantum releases about 300 electrons. The X-ray sensitivity is too low for medical applications.

2242: F. H. J. van der Poel: A 35 mm film scanner (T. Ned. Radiogenootsch. **20**, 11-19, 1955, No. 1).

After some general remarks on the televising of motion-picture film and a brief description of a well-known film scanning method using an intermittent projector and a storage type camera tube, this paper describes a method using the flying spot scanner (see Philips tech. Rev. **15**, 221-232, 1953/1954) and continuously moving film. Television pictures of very good quality are obtained with the latter method.

2243: K. Teer: Overdrachtsystemen voor kleuren-televisie (T. Ned. Radiogenootsch. **20**, 21-34, 1955, No. 1). (Transmission systems for colour television; in Dutch).

A compatible colour television system is required to transmit the colour picture information, which can be represented by three video signals, in a normal television channel in such a way that monochrome reception on existing black and white receivers is possible. Such a transmission can be established by adding to the normal video signal (luminance signal) subcarriers modulated by two other signals (colour signals). Two subcarriers may be used for the transmission of the colour signals but the modulation of both signals on a single subcarrier is also possible. (N.T.S.C. colour system in U.S.A.) A brief description is given of an experimental transmisson system where two subcarriers are applied.

2244: A. G. van Doorn and F. W. de Vrijer: Display methods for colour television (T. Ned. Radiogenootsch. **20**, 35-48, 1955, No. 1).

This article describes some possible methods of producing colour-television images when three simultaneous colour signals representing the red, green and blue brightnesses are available. In addition to a brief review of the principles used in a direct-view colour-picture tubes, a description is given of a projection method of displaying colour images. The results obtained to date by the various systems are compared.

2245: F. A. Kröger: The physical chemistry of sulphide phosphors (Brit. J. appl. Phys., Suppl. No. 4, 558-564, 1964).

By preparing phosphors under controlled conditions, it is possible to regulate the properties in a reproducible way. From a study of the variations induced in this way, insight into the micro-structure of the phosphor can be obtained. For ZnS, various physico-chemical aspects of the preparation such as oxidation-reduction, the action of fluxes and the formation of solid solutions are discussed.